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An X-Y Plotter Stabilising Adaptor for use with Sampling Oscilloscopes

B. Gilbert
Mullard Central Application Laboratory

The equipment described enables an X-Y plotting table to be used with a sampling oscilloscope with the maximum amount of operational simplicity. Scanning rate is adjusted automatically in a way which depends on the waveform which is being displayed, so that the graph is produced in the minimum possible time without distortion due to carriage overshoot or lag. Several other features are incorporated which ensure virtually foolproof operation; these include automatic pen-lift under overload conditions, calibrated scale multipliers, and automatic transfer from dynamic (c.r.t.) display to the plotting condition.

INTRODUCTION

As is well known, a waveform displayed on a sampling oscilloscope may be scanned at low speed and traced out on graph paper by means of an X-Y plotting table. This facility is very useful, but tends not to be practised very much. The reasons for this, apart from the high cost of X-Y plotters, are thought to be as follows:

(i) Interconnection is often tedious.

(ii) Calibration of scales must usually be performed each time the plotter is reconnected.

(iii) The transfer from the dynamic (c.r.t.) display to the plotting condition must be done manually for each graph, including switching the scanning mode and pen-lift controls.

(iv) Since the plotter remains connected whilst the waveform is displayed on the c.r.t., the carriage is disturbed erratically, often causing ink to be splashed onto the paper.

(v) Incorrect traces can be made if the oscilloscope Y-amplifiers limit or if the carriage travel is limited in either direction.

The equipment described was designed to simplify the use of a graph-plotter. With this equipment, it is only necessary to plug in to the oscilloscope, place a sheet of paper on the plotting table, select the required display axes and scale multiplication factors and press the 'Start' button. The transfer from scanning by the internally-generated staircase of the oscilloscope to a slow ramp, and pen-lowering, are automatic. The rate of scanning is continually adjusted automatically to keep the resolved carriage speed (scalar quantity) constant, thus permitting the waveform to be traced in the minimum possible time for a given table and obviating the need to select the scanning rate. Any overload condition, which might distort the waveform, is indicated by a warning light and causes the pen to be lifted. At the end of the cycle, the pen is lifted, the carriage is returned to its starting position, and the c.r.t. display reappears. Repetitive operation can be used if required. Other miscellaneous features of this equipment are:

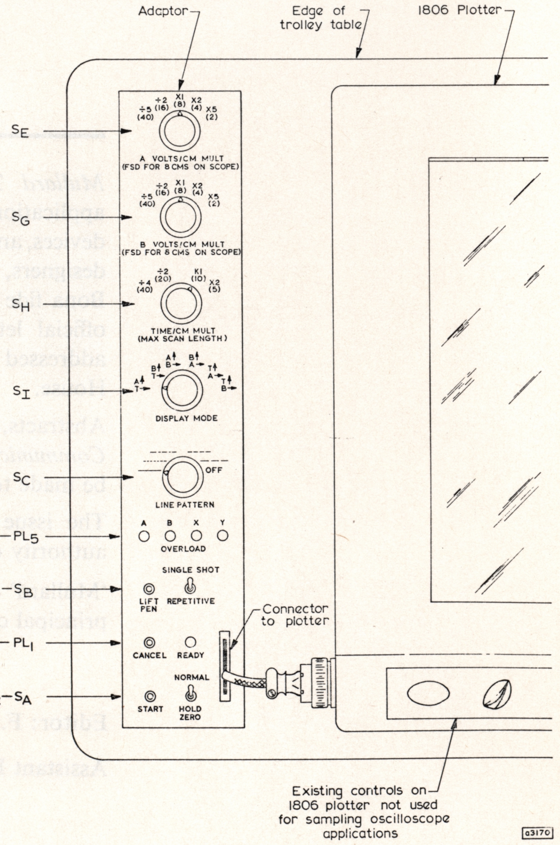

Adaptor
Edge of trolley table
1806 Plotter
S_E
÷5 (40)
÷2 (16)
X1 (8)
X2 (4)
X5 (2)
A VOLTS/CM MULT (FSD FOR 8 CMS ON SCOPE)
S_G
B VOLTS/CM MULT (FSD FOR 8 CMS ON SCOPE)
S_H
÷4 (40)
÷2 (20)
X1 (10)
X2 (5)
TIME/CM MULT (MAX SCAN LENGTH)
S_I
DISPLAY MODE
S_C
OFF
LINE PATTERN
PL_2-PL_5
A B X Y
OVERLOAD
PB_3-S_B
SINGLE SHOT
LIFT PEN
REPETITIVE
PB_1-PL_1
CANCEL
READY
PB_2-S_A
NORMAL
START
HOLD ZERO
Connector to plotter
Existing controls on 1806 plotter not used for sampling oscilloscope applications
a3170

Fig. 1—Frontal layout of prototype

(i) The cycle may be cancelled at any time.

(ii) The pen may be lifted at any time.

(iii) Various line structures can be used (continuous, dotted, dashed, etc.).

(iv) The following axes can be used:

Vertical		*Horizontal*
±A	versus	Time
±B	versus	Time
±A	versus	±B
±B	versus	±A
Time	versus	±A
Time	versus	±B

Also, by using the 'A+B' output, available on most sampling oscilloscopes, ±A±B can be plotted versus time with a choice of orthogonal axes.

(v) A-axis or B-axis multiplying factors of ÷5, ÷2, ×1, ×2, ×5 are available.

(vi) Time-axis multiplying factors of ÷4, ÷2, ×1, ×2 (giving sweep lengths of 40cm, 20cm, 10cm, 5cm) are available.

(vii) The starting-point of the sweep can be held, to enable accurate positioning of the trace.

(viii) Filters can be used to reduce noise level. With the present equipment, sensitivities up to 400μV/cm are possible.

This equipment was designed specifically for use with a Tektronix 561 oscilloscope fitted with 3T77 and 3S76 plug-in units (although a Tektronix 661 is also suitable), and a Bryans Aeroquipment Model 1806 Plotting Table; small modifications are desirable to both. Very little modification is necessary to adapt the equipment to other combinations. The complete prototype equipment was mounted together with the plotting table on a trolley, for convenience of portability. Panel controls are shown in Fig. 1.

STABILISING THE PEN SPEED

Initially, we will consider only the c.r.t. display, with the aid of Fig. 2. If the slope of the waveform, as displayed on particular time and voltage scales, at any position along the x-axis is $\gamma(x)$, and the horizontal scanning velocity is $\dot{x}$ the vertical rate of motion $\dot{y}$ is given by

$$\dot{y} = \gamma(x)\dot{x}. \qquad \ldots(1)$$

The spot speed S is therefore given by

$$\begin{aligned} S(x) &= \sqrt{(\dot{x}^2+\dot{y}^2)} \\ &= \dot{x}\sqrt{(1+[\gamma(x)]^2)}. \qquad \ldots(2) \end{aligned}$$

Ideally, we desire S to be constant (so that the spot or carriage is always moving at maximum speed). This requires that

$$x(t) = \int_0^t \frac{S}{\sqrt{(1+[\gamma(x)]^2)}}dt.$$

Now, to solve this equation electrically would require, in addition to an integrator, the means for squaring, square-rooting and dividing. Also, the actual (geometric) waveform slope, $\gamma(x)$, would have to be continuously measured.

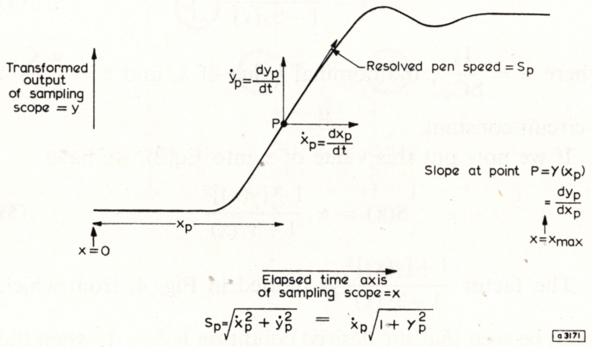

Transformed output of sampling scope = y
$\dot{y}_p=\frac{dy_p}{dt}$
Resolved pen speed = S_p
P
$\dot{x}_p=\frac{dx_p}{dt}$
Slope at point $P=\gamma(x_p)$
$=\frac{dy_p}{dx_p}$
x_p
x=0
$x=x_{max}$
Elapsed time axis of sampling scope=x
$S_p=\sqrt{\dot{x}_p^2+\dot{y}_p^2} = \dot{x}_p\sqrt{1+\gamma_p^2}$
a3171

Fig. 2—Basic vectors

A simpler scheme is shown in Fig. 3. This only uses a simple differentiator, integrator and subtraction circuit, plus a circuit for taking the modulus of the slope.

V_1 represents the y-voltage, and this is related to actual y position on the c.r.t. through the factor 'a' (V/cm). The current in C_1 is

$$I_1 = C_1.\frac{dV_1}{dt} = aC_1\dot{y}.$$

I_2 has the same magnitude, but is always positive. This current is subtracted from the control current, I_3,

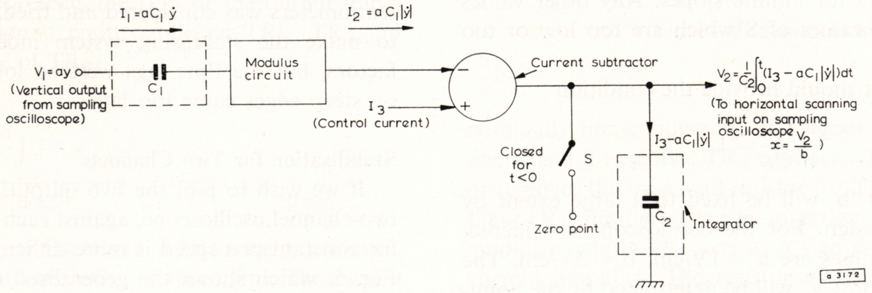

$I_1=aC_1\dot{y}$
$I_2=aC_1|\dot{y}|$
$V_1=ay$
(Vertical output from sampling oscilloscope)
C_1
Modulus circuit
I_3
(Control current)
−
+
Current subtractor
$V_2=\frac{1}{C_2}\int_0^t(I_3-aC_1|\dot{y}|)dt$
(To horizontal scanning input on sampling oscilloscope $x=\frac{V_2}{b}$)
$I_3-aC_1|\dot{y}|$
Closed for t<0
S
Zero point
C_2
Integrator
a 3172

Fig. 3—Stabiliser arrangements

and the difference is integrated in C_2. The voltage across C_2 is reset to zero before each cycle. Thus,

$$V_2 = \frac{1}{C_2}\int_0^t (I_3 - aC_1\dot{y})dt.$$

The x position on the c.r.t. is related to V_2 through the constant 'b' (V/cm). Thus we may write

$$\dot{x} = \frac{d}{dt}\cdot\frac{1}{b.C_2}\int_0^t (I_3 - aC_1\dot{y})dt$$

$$= \frac{I_3}{bC_2} - \frac{a}{b}\cdot\frac{C_1}{C_2}\dot{y}. \qquad \ldots(3)$$

But Eq(1) states that $\dot{y} = \gamma(x)\dot{x}$ so that

$$\dot{x}\left[1 + \frac{a}{b}\cdot\frac{C_1}{C_2}\gamma(x)\right] = \frac{I_3}{bC_2}$$

or

$$\dot{x} = \frac{\alpha}{1+\beta\gamma(x)} \qquad \ldots(4)$$

where $\alpha = \frac{I_3}{bC_2}$, the nominal value of $\dot{x}$, and $\beta = \frac{\alpha}{b}\cdot\frac{C_1}{C_2}$, a circuit constant.

If we now put this value of $\dot{x}$ into Eq(2), we have

$$S(x) = \alpha.\frac{1+[\gamma(x)]^2}{1+\beta\gamma(x)}. \qquad \ldots(5)$$

The factor $\frac{1+[\gamma(x)]^2}{1+\beta\gamma(x)}$ is plotted in Fig. 4, from which it can be seen that the desired condition is $\beta = 1$, when the pen speed reduces to 70·7% of α for $\gamma(x) = 1$, and rises again to 100% of α for infinite slopes. Any other values of β give terminal values of S which are too low or too high.

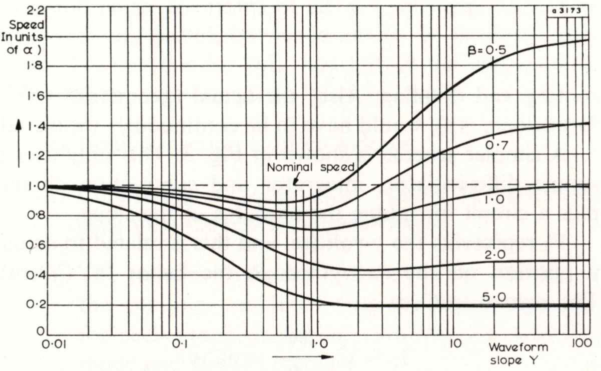


Fig. 4—Plotting speed versus γ

Thus, the circuit should provide the condition

$$\frac{a}{b}\cdot\frac{C_1}{C_2} = 1.$$

In practice 'a' and 'b' will be fixed to a large extent by the oscilloscope design. For the oscilloscope considered, the approximate values are a = 1V/cm, b = 5V/cm. The integrating capacitor, C_2, will be determined by the nominal value of I_3, and the nominal speed desired. If these are 200μA and 1cm/s respectively, C_2 must be 40μF. It follows that

$$C_1 = \frac{bC_2}{a} = 200\mu F.$$

So far, we have assumed that the plotter carriage would follow the same movement as the spot. As mentioned earlier, however, multiplication factors can be introduced to alter the size of the trace. If the x and y multiplication factors are M_x and M_y, the condition which must now be satisfied is

$$\frac{M_y a C_1}{M_x b C_2} = 1.$$

Additionally, if α is to remain unaffected, we must alter I_3 to a new value equal to the old value divided by M_x.

This makes for difficult, albeit soluble, switching problems, and it was decided to design the system for scale factors of ×2 in both horizontal and vertical directions; that is, the c.r.t. area of 18cm×10cm is increased to a scanning area of 36cm×20cm; the 36cm assumes the scope amplifiers overload at ±9cm (see page 99). If larger traces are desired, pen speeds are increased, but not seriously, whilst if smaller traces are made, plotting time is longer than it need be. The nominal plotting time for unmodulated line is about fifteen seconds, but may rise to many times this for complex waveforms. As a matter of interest, operation of the stabilising system direct from the x- and y-position voltages obtained from the feedback potentiometers was considered and tried, the intention being to make the stabilising system independent of scaling factors, but the time lags were so long that stabilisation on steep edges came too late.

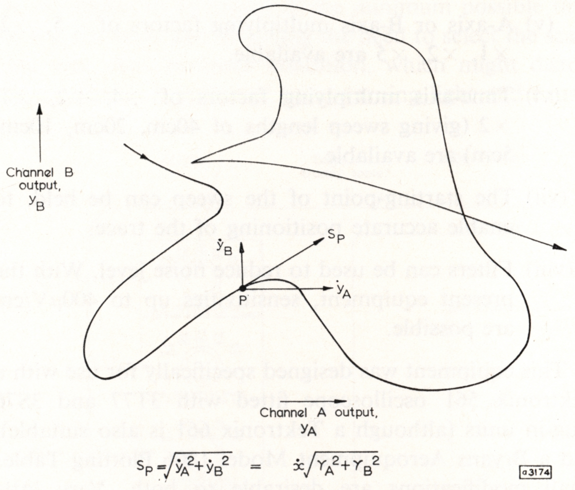

Channel B output, y_B
$\dot{y}_B$
S_P
$\dot{y}_A$
P
Channel A output, y_A
$S_P = \sqrt{\dot{y}_A{}^2 + \dot{y}_B{}^2} = \dot{x}\sqrt{\gamma_A{}^2 + \gamma_B{}^2}$
a3174

Fig. 5—Vectors for two-channel case

Stabilisation for Two Channels

If we wish to plot the two outputs, say A and B, of a two-channel oscilloscope, against each other, the condition for constant pen speed is more difficult to meet. Consider Fig. 5, which shows the generalised case. It is seen that

$$S(x) = \sqrt{\dot{y}_A{}^2 + \dot{y}_B{}^2} = \dot{x}\sqrt{\gamma_A(x)^2 + \gamma_B(x)^2}.$$

Thus, the x position should be given by

$$x(t) = \int_0^t \frac{S}{\sqrt{[\gamma_A(x)^2 + \gamma_B(x)^2]}}.dt,$$

which, for solution, again requires undesirable complexity in the stabilising circuit.

By simply adding the moduli of the differential currents proportional to $\dot{y}_A$ and $\dot{y}_B$ the term $a.C_1.\dot{y}$ becomes $a.C_1.(\dot{y}_A + \dot{y}_B)$, if the differentiators are identical, or $a.C_1.\dot{x}[\gamma_A(x) + \gamma_B(x)]$, so that Eq(4) becomes

$$\dot{x} = \frac{\alpha}{1+\beta[\gamma_A(x)+\gamma_B(x)]},$$

and Eq(5) becomes

$$S(x) = \alpha.\frac{\gamma_A(x)^2 + \gamma_B(x)^2}{1+\beta[\gamma_A(x)+\gamma_B(x)]}.$$

Fig. 6 shows how S varies with γ_A and γ_B for $\beta = 1$, which is the only case of interest. It can be seen that the condition of constant pen speed is not met, but that when either of the slopes is greater than about unity, it is always above 50% of the nominal, and never exceeds 100%. This method is therefore considered adequate. (See also *Performance*, page 101).

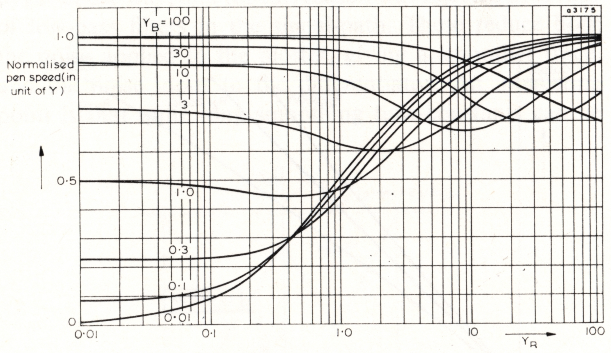


Fig. 6—Two-channel curves

Basic Differentiator

It is not essential to use very accurate circuits for the stabilising system. A simple differentiator is shown in Fig. 7a, where TR_1 and TR_2 form a double emitter-follower giving a low output impedance, and TR_3 forms a low input impedance stage. Provided that the current I_1, which equals $C_1.dV/dt$, is small compared to the tail currents, I, the equivalent circuit of Fig. 7b can be used. The resistance r_o represents the sum of the output impedance of the compound emitter-follower TR_1, TR_2 and the input impedance of TR_3.

That is,

$$r_o = \frac{\left(\frac{R_s + r_{b1}}{h_{FE1}}\right) + r_{b2} + r_{e1}}{h_{FE2}} + r_{e2} + r_{e3} + \frac{r_{b3}}{h_{FE3}}.$$

In practice, a value of r_o less than 25Ω is easily achieved, giving a time-constant, with C_1, of about 5ms. The current I_1 reaches its correct value after about 3 or 4 time-constants, that is, within 20ms. Even assuming no stabilising action during this time, the carriage cannot move very far, because of its inertia, before full stabilisation comes in; in other words, this approximation to a true differentiator is adequate.

Basic Modulus Circuit

The modulus circuit produces an output current equal to the modulus of the differential current through C_1. The basic arrangement is shown in Fig. 8. The differential current I_1 in the collector of TR_3 produces variations in e_1. The steady collector current of TR_3, $\alpha_3 I_2$, is backed off by the current in R_1, so that both TR_4 and TR_5 are nominally non-conducting. If e_1 goes positive, TR_4 conducts, and if negative, TR_5 conducts. The diode D inserts or removes the base load resistor for TR_4. It is easily seen that TR_4 produces a sign-inverting unity-gain current amplifier, whilst TR_5 acts as a sign-preserving unity-gain current amplifier. The resistor in the emitter of TR_5 is included only to keep e_1 nearly proportional to I_1.

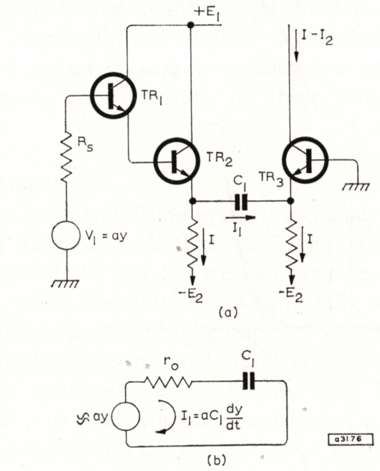


Fig. 7—Basic differentiator *(a)* and simple equivalent *(b)*

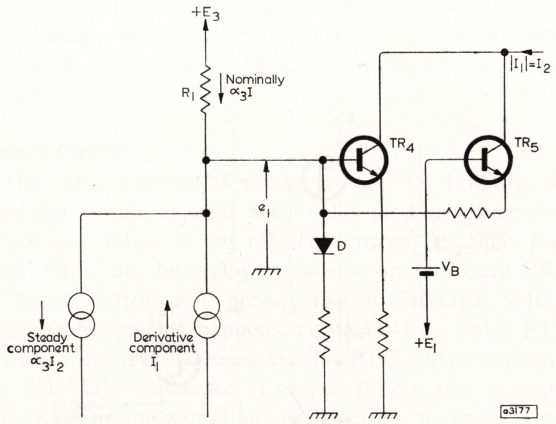


Fig. 8—Modulus circuit

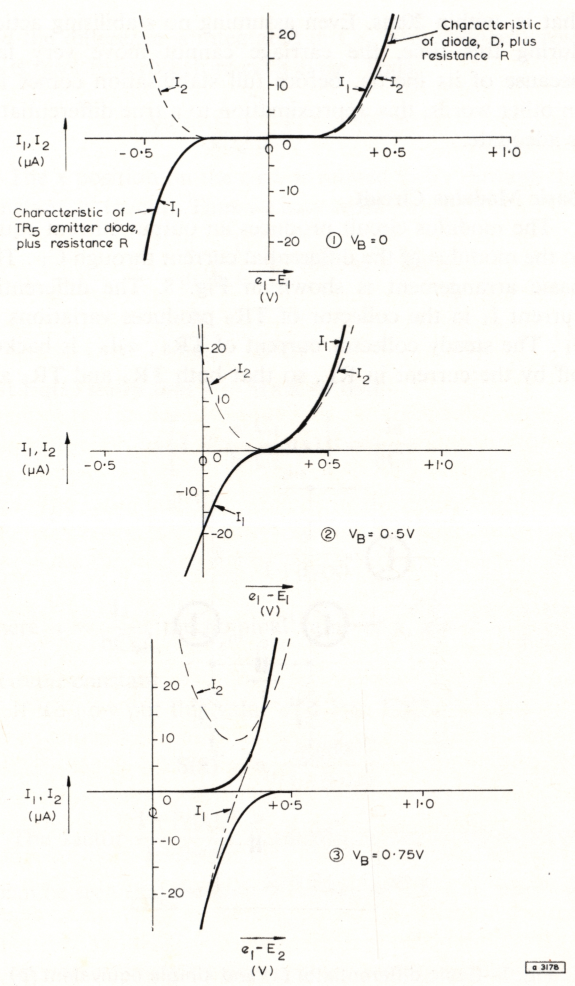


Fig. 9—Effect of V_B on e_1/I_1 curves

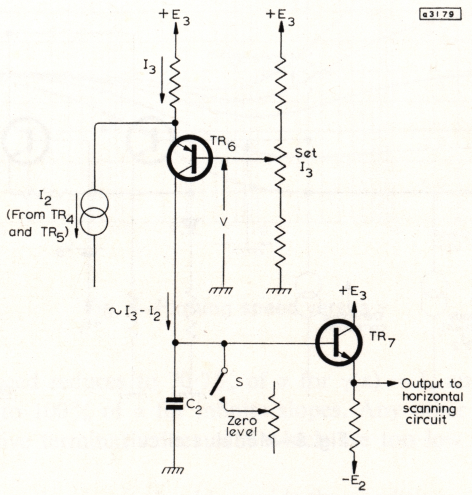


Fig. 10—Subtraction and integration circuit

The bias V_B is included to avoid a dead-zone. Fig. 9 shows how the e_1/I_2 characteristics near the origin are affected by V_B. A value of V_B slightly too high is allowable, since this only introduces a constant term in I_2, which can be taken into account by increasing the integrator control current.

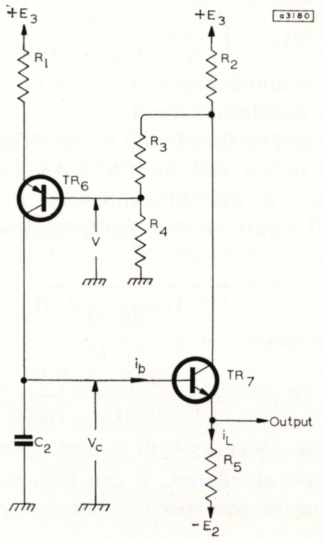


Fig. 11—Method of making the integrator linear.

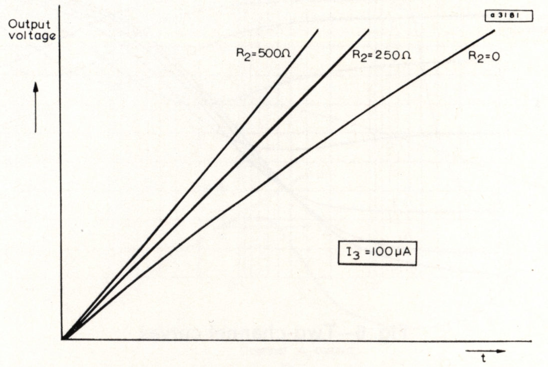


Fig. 12—Integrator output with linearising circuit

Basic Subtraction Circuit and Integrator

The circuit used to subtract the current I_2 from the control current and integrate the difference is shown in basic form in Fig. 10, and is more or less self-explanatory. C_2 normally charges at a rate proportional to I_3. Any change in the Y position produces a value of I_2 proportional to the rate of this change and reduces the charging rate. Before each cycle starts, S is closed to define the integrator zero-level. The maximum positive voltage is limited by V. The emitter-follower provides a reasonable shunt resistance across C_2. This can be raised to near-infinite values by the use of the feedback circuit shown in Fig. 11. As V_c rises, the base current, i_b, in TR_7 also rises, due to the increase in the low current i_L. However, the voltage V is also reduced by the presence of the feedback resistor R_2, which increases I_3 to offset the fall in charging current. It can be easily shown that for compensation,

$$R_2 \simeq \frac{R_1(R_3+R_4)}{R_4 h_{FE7}}.$$

The linearity of the practical integrator can be judged from Fig. 12, which shows the effect of R_2. It can also be shown that the critical value of R_2 is sensibly unaffected by the load resistor R_5.

Operation of System in the Presence of Noise

The analysis given for the system assumes a noise-free signal. In practice, noise will be caused by

(i) the 'steady-state' noise of the sampling channel. (This increases with increasing sensitivity setting.)
(ii) the jitter-generated noise. (This increases with increasing time-scale as well as sensitivity setting.)

The effect of the former component is to produce a steady current output from the differentiator, and cause the rate of horizontal progression to be reduced by a constant factor; this factor increases on the more sensitive ranges. This effect could be mitigated by increasing the differentiator time-constant. The output of a perfect differentiator to white noise spectrum is a linearly rising spectrum. The mean output current is proportional to the area under the output spectrum curve. If now the differentiator is given a finite time-constant, the output spectrum flattens off at high frequencies with a corresponding reduction in the mean output current. This benefit is achieved at the expense of the time taken for stabilisation to get under way, and for this reason was rejected in favour of low-pass filters in the signal-path. These reduce both the noise applied to the plotter, giving a smoother trace, and the noise input to the differentiator. An interesting point is that quite long smoothing time-constants can be used because the action of the stabiliser on the transformed waveform is to remove all the high-frequency components.

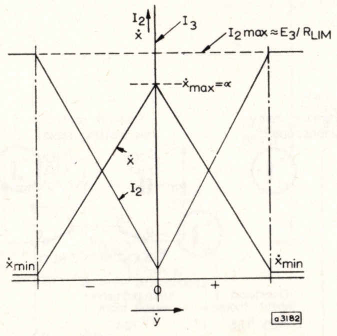


Fig. 13—Modified control characteristic

The noise due to the second cause, jitter, is rather more serious. When displaying fast-rise pulses on long time-scales, the indecision in sampling time as the edge is scanned can cause noise voltages very nearly equal to the pulse amplitude. Even with the measures outlined above, the output from the differentiator can reach such high values that the integrator subtraction circuit is completely swamped (that is, $I_2 > I_3$ in Fig. 10), and horizontal progression ceases. This, in turn, causes the noise voltage to remain, because the oscilloscope will not move off the edge, and the system grinds to a halt.

This problem can be overcome to a degree sufficient for all but the most difficult waveforms, by limiting the output current from the differentiator to a value less than I_3 by including a resistor R_{LIM} in series with the collectors of TR_4 and TR_5. This produces the control characteristic shown in Fig. 13. Scanning continues very slowly (at a rate $\dot{x}_{min}$) even under complete overload, so that the noisy edge is eventually passed. The loss of control during this time may lead to carriage speeds greater than desired, and a better method of dealing with jitter-generated noise is still sought. The low-pass filters, of course, can also be used to advantage.

COMPLETE SYSTEM

The complete equipment uses a total of forty-three transistors and eight relays. Few transistor types, and only one diode type, are used, and these are all silicon devices. The power supply is conventional and is not described. The power required is +64V at 30mA, −12V at 40mA and either +12V at 550mA or +12V at 160mA, plus an a.c. supply for the indicator lamps of about 12V at 400mA. The d.c. supplies must be stable over long periods; some of the circuits (for example, the modulus circuit) are critically dependent on supply rail stability, although the initial values of supply voltages are not critical. In this respect, improvements to these circuits may be desirable.

Block Diagram

To the basic stabilising system, there have been added (Fig. 14) the scale multipliers (1, 2 and 3), the necessary control circuits (4 and 5), overload sensing circuits (6, 7, 8 and 9), and a line-pattern generator (10), which lifts the pen rapidly to produce broken lines of various patterns.

The display selector switch S_9 also selects the conditions for which overloads should lift the pen. For example, if channel A is being plotted versus time, an overload on channel A should lift the pen, whilst an overload on channel B is unimportant; whereas when A is plotted versus B, an overload in either should lift the pen. X- and Y-overloads on the table itself always lift the pen. These are sensed by measuring the servo feedback voltages, made available by a modification to the plotting table (see Appendix 1).

Similarly, the integrator is controlled from either channel A, channel B or both, depending on the chosen display.

Control Circuit

The control circuit is shown in Fig. 15. Initially, the integrator output is near zero volts, so D_{601} is reverse-biased and TR_{605} is cut off. The current through R_{613} holds TR_{701} on, but relay A cannot operate until PB_2, the 'Start' button, is depressed (S_B in 'SINGLE SHOT' position). When this happens, contact RLA_1 holds RLA on while contact RLA_4 releases relay RLC and extinguishes the 'READY' indicator. Contact RLA_4 also operates relay RLJ in the sampling oscilloscope, to transfer the scanning control to the integrator. Contact RLC_2 closes, and contact RLC_1 switches the base of TR_{703} to a negative potential through R_{705}, so that after a short delay due to C_{702}, introduced to allow the carriage to reach its correct starting point, relay RLD operates, closing RLD_1 and lowering the pen. The operation of RLD can be inhibited, if

(i) an overload state exists,
(ii) the 'LIFT PEN' button, PB3, is depressed,
(iii) $SW1_1$ is in the 'HOLD ZERO' position, or
(iv) the line pattern generator is inhibiting.

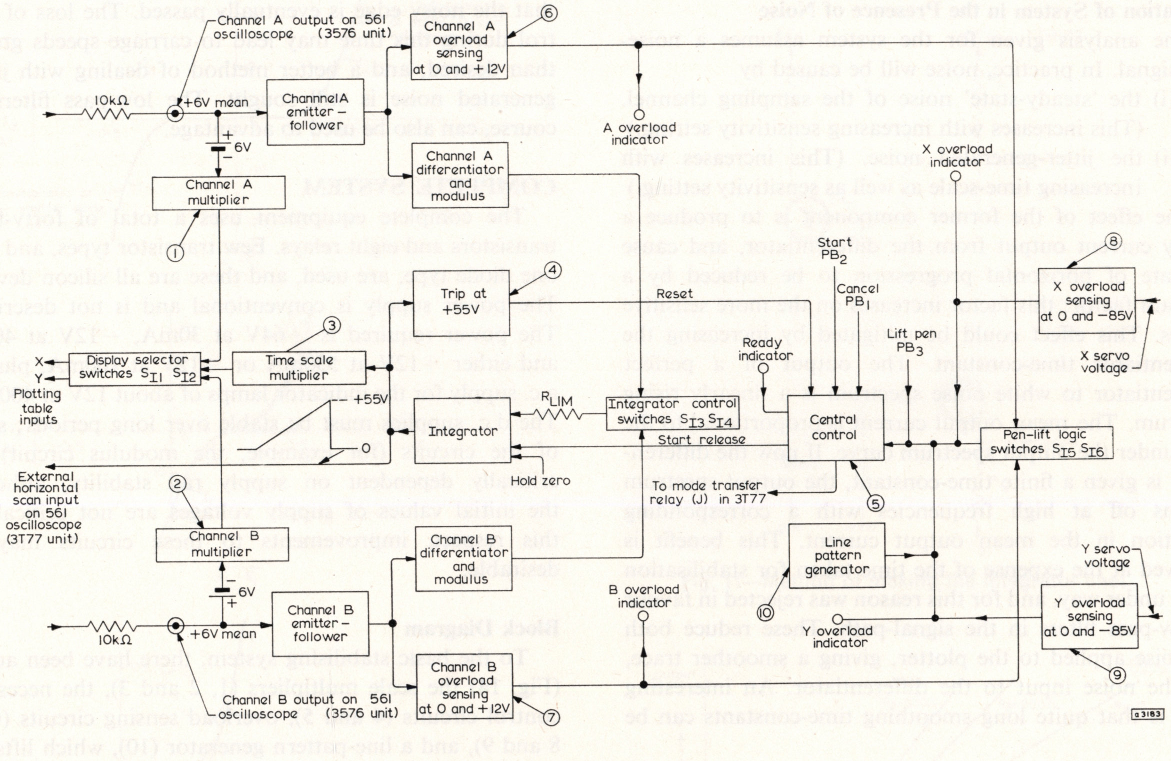
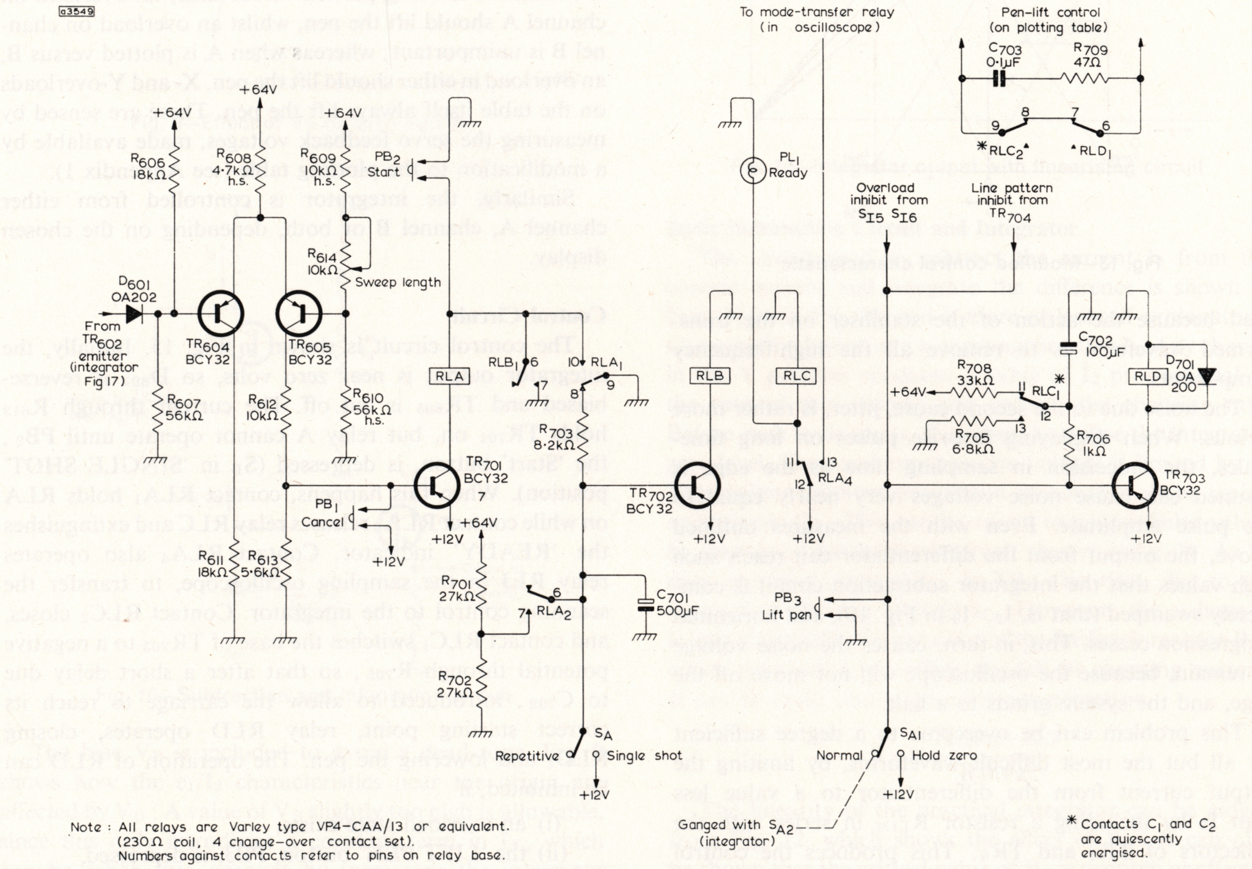

Channel A output on 561 oscilloscope (3576 unit)
Channel A overload sensing at 0 and +12V
Channel A emitter-follower
Channel A differentiator and modulus
Channel A multiplier
10kΩ
+6V mean
6V
A overload indicator
X overload indicator
Trip at +55V
Reset
Start PB2
Cancel PB1
Lift pen PB3
Ready indicator
X overload sensing at 0 and −85V
X servo voltage
Display selector switches SI1 SI2
Time scale multiplier
Plotting table inputs
+55V
RLIM
Integrator
Integrator control switches SI3 SI4
Start release
Central control
Pen-lift logic switches SI5 SI6
Hold zero
External horizontal scan input on 561 oscilloscope (3T77 unit)
To mode transfer relay (J) in 3T77
Channel B multiplier
Channel B differentiator and modulus
Line pattern generator
Y servo voltage
B overload indicator
Y overload indicator
Y overload sensing at 0 and −85V
Channel B emitter-follower
Channel B overload sensing at 0 and +12V
Channel B output on 561 oscilloscope (3576 unit)
A3183

Fig. 14—Block diagram of system


To mode-transfer relay (in oscilloscope)
Pen-lift control (on plotting table)
C703 0·1μF
R709 47Ω
RLC2
*RLD1
+64V
R606 18kΩ
R608 4·7kΩ h.s.
R609 10kΩ h.s.
PB2 Start
PL1 Ready
Overload inhibit from SI5 SI6
Line pattern inhibit from TR704
R614 10kΩ
Sweep length
D601 OA202
From TR602 emitter (integrator Fig 17)
TR604 BCY32
TR605 BCY32
RLA
RLB1
RLA1
RLB
RLC
RLD
C702 100μF
D701 OA200
R607 56kΩ
R612 10kΩ
R610 56kΩ h.s.
R703 8·2kΩ
R708 33kΩ
RLC1
R705 6·8kΩ
R706 1kΩ
TR701 BCY32
TR702 BCY32
TR703 BCY32
RLA4
PB1 Cancel
+12V
R611 18kΩ
R613 5·6kΩ
R701 27kΩ
RLA2
C701 500μF
PB3 Lift pen
R702 27kΩ
SA
Repetitive
Single shot
SA1
Normal
Hold zero
Ganged with SA2 (integrator)

Note: All relays are Varley type VP4-CAA/13 or equivalent. (230 Ω coil, 4 change-over contact set). Numbers against contacts refer to pins on relay base.

* Contacts C_1 and C_2 are quiescently energised.

Fig. 15—Control circuits

The 'zero' position of the carriage can be held indefinitely by S_A, which clamps the integrating capacitor C_{704} (Fig. 16) to the starting level, as well as keeping the pen raised.

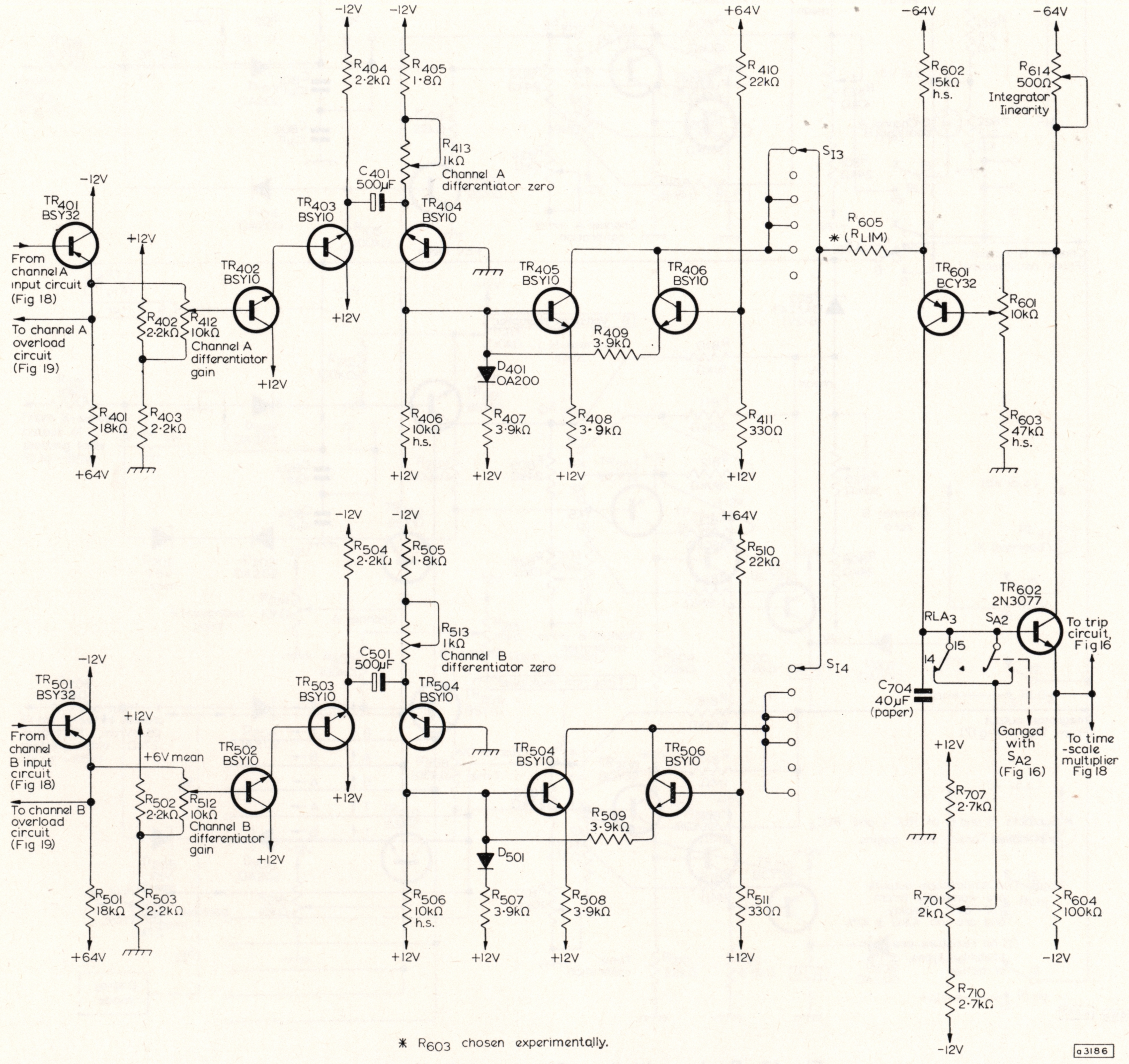
When the integrator output reaches the trip level defined by R_{609}, R_{614} and R_{610}, TR_{605} conducts and switches off TR_{701} to release RLA. The pen is then lifted, the integrator is reset and the oscilloscope reverts to a continuous display. The re-illumination of the 'READY' light indicates that another cycle may be started. The cycle may be cancelled at any time by PB_1.

Repetitive operation is provided for by TR_{702} and relay RLB. If S_B is put into the 'REPETITIVE' position, R_{703} turns TR_{702} on via contact RLA_1, thus operating RLB which momentarily closes contact RLB_1 to operate RLA, after which the action is as before. Shortly after RLA_1 changes over, the base of TR_{702} rises to +32V, so that when RLA releases at the end of the cycle there is a delay of about 3s before the next cycle is initiated, allowing ample time for the carriage to return and giving a brief display on the waveform on the c.r.t.

Stabilising Circuits

The complete stabilising circuits for two channels are shown in Fig. 16 (circuit references 400+ for channel A, 500+ for channel B). Since the operation has been discussed above, a detailed account is unnecessary. The current corresponding to I_2 of Fig. 6 is adjusted by means of R_{413} or R_{513} to give a minimum standing current (about 5μA) in the collectors of $TR_{405}+TR_{406}$ or $TR_{504}+TR_{505}$. The differentiator gain controls R_{412}, R_{512} can be adjusted on test waveforms to give the desired condition of $\beta = 1$.

S_{13} and S_{14} select the outputs from the modulus circuits, according to the display mode. The sum of these currents is limited by R_{601} to a little less than 150μA. The nominal integrator current is set up by means of R_{601} to 150μA, giving a nominal ramp time of 15 seconds. This adjustment is best finalised by timing the uncontrolled ramp. Ramp linearity can be adjusted by means of R_{614}, using the line pattern generator (described below) to produce time markers.


−12V
+64V
−64V
R404 2·2kΩ
R405 1·8Ω
R410 22kΩ
R602 15kΩ h.s.
R614 500Ω Integrator linearity
R413 1kΩ Channel A differentiator zero
C401 500μF
TR401 BSY32
TR403 BSY10
TR404 BSY10
S13
R605 *(RLIM)
From channel A input circuit (Fig 18)
TR402 BSY10
TR405 BSY10
TR406 BSY10
TR601 BCY32
R402 2·2kΩ
R412 10kΩ Channel A differentiator gain
To channel A overload circuit (Fig 19)
R409 3·9kΩ
R601 10kΩ
D401 OA200
R401 18kΩ
R403 2·2kΩ
R406 10kΩ h.s.
R407 3·9kΩ
R408 3·9kΩ
R411 330Ω
R603 47kΩ h.s.
+12V
R504 2·2kΩ
R505 1·8kΩ
R510 22kΩ
TR602 2N3077
R513 1kΩ Channel B differentiator zero
C501 500μF
RLA3
SA2
To trip circuit Fig 16
S14
TR501 BSY32
TR503 BSY10
TR504 BSY10
C704 40μF (paper)
From channel B input circuit (Fig 18)
+6V mean
TR502 BSY10
TR504 BSY10
TR506 BSY10
Ganged with SA2 (Fig 16)
To time-scale multiplier Fig 18
R502 2·2kΩ
R512 10kΩ Channel B differentiator gain
To channel B overload circuit (Fig 19)
R509 3·9kΩ
R707 2·7kΩ
D501
R501 18kΩ
R503 2·2kΩ
R506 10kΩ h.s.
R507 3·9kΩ
R508 3·9kΩ
R511 330Ω
R701 2kΩ
R604 100kΩ
R710 2·7kΩ
A3186

* R_{603} chosen experimentally.

Fig. 16—Two-channel stabiliser

Scale Multipliers and Display-mode Switching

Fig. 17 shows the attenuators used with the Bryans table type 1806. They are based on the assumption that the input impedance of the X and Y channels is 27kΩ, and the sensitivity is 25mV/cm. This input impedance in parallel with the combinations $R_{802}+R_{824}$, $R_{809}+R_{826}$, and $R_{820}+R_{827}$ produces a standardised value of about 12·5kΩ for the lower resistor of each attenuator.

The current through $R_{823}+R_{801}+D_{801}$ shifts the 'A' output of the oscilloscope by −6·2V, and this same current flowing in the 10kΩ output impedance of the oscilloscope sets the level at the scope output point to +6·2V, when the scope trace is centred. Thus, the input to the plotter for a centred trace is zero, when R_{823} is correctly adjusted.

During the 'READY' state, relay RLC is energised, so that the input to the system is taken to a fixed +6V by contact RLC_3. This prevents the differentiator overloading and keeps the carriage stationary.

It is important if modifications are made to ensure that this quiescent value of +6V be maintained, since it affects the d.c. level in the differentiator, and in the overload detectors.

Low-pass filtering is available by means of S_D and C_{801}, C_{802}, giving time-constants of about 35, 70 and 140ms, the first value being permanently in circuit.

Channel B circuits are identical.

S_{I1} and S_{I2} select channel A, or B, or the integrator output applying them to the X and Y inputs as required. It is important that X and Y sensitivities are equal.

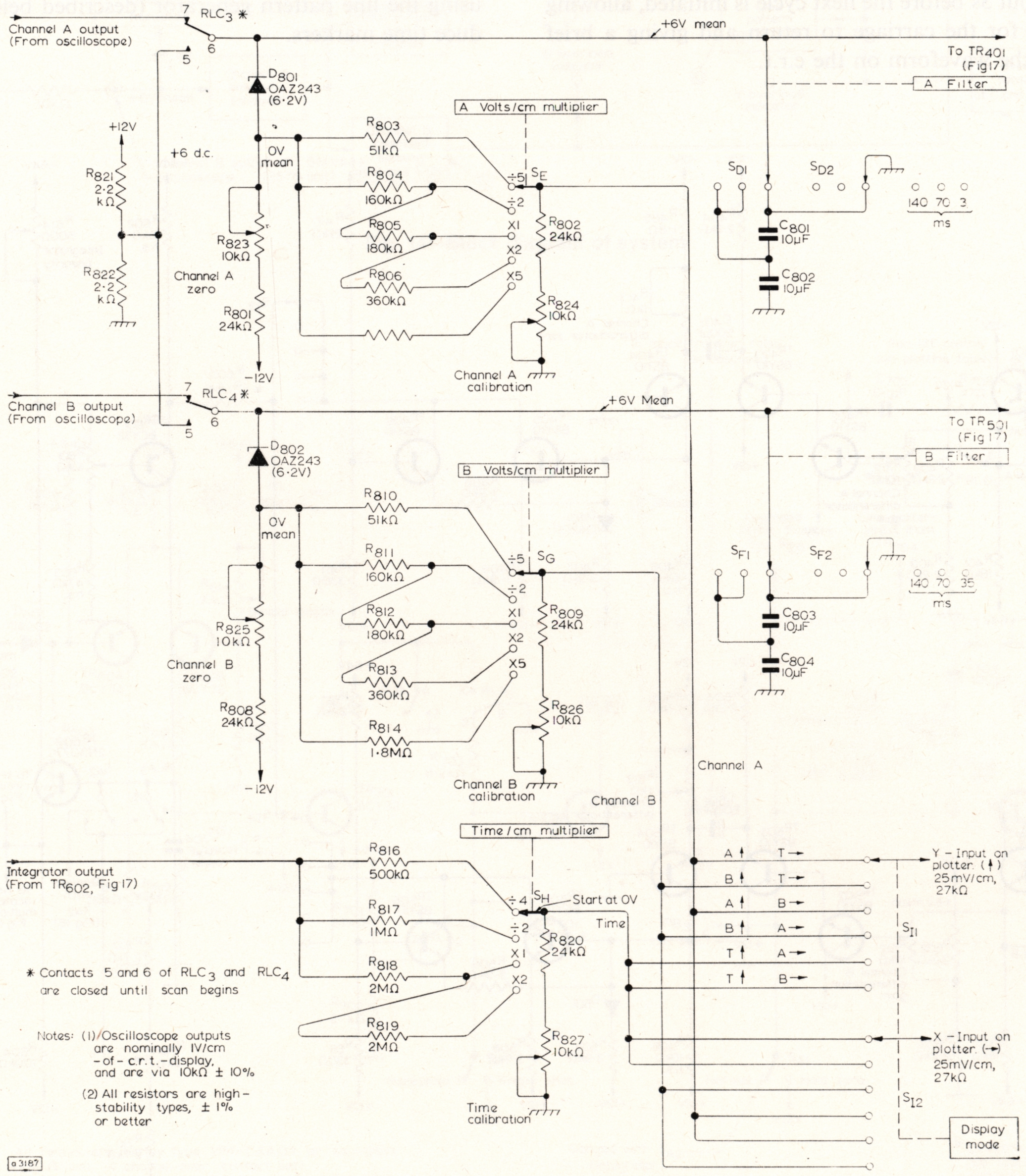

Channel A output
(From oscilloscope)
7
RLC3 *
6
5
+6V mean
To TR401
(Fig17)
A Filter
D801
OAZ243
(6·2V)
A Volts/cm multiplier
+12V
R821
2·2
kΩ
+6 d.c.
0V
mean
R803
51kΩ
R804
160kΩ
R805
180kΩ
R806
360kΩ
÷5
SE
÷2
X1
X2
X5
R802
24kΩ
R824
10kΩ
R823
10kΩ
R822
2·2
kΩ
Channel A
zero
R801
24kΩ
−12V
Channel A
calibration
SD1
SD2
140 70 3
ms
C801
10µF
C802
10µF
Channel B output
(From oscilloscope)
7
RLC4 *
6
5
+6V Mean
To TR501
(Fig 17)
B Filter
D802
OAZ243
(6·2V)
B Volts/cm multiplier
0V
mean
R810
51kΩ
R811
160kΩ
R812
180kΩ
R813
360kΩ
R814
1·8MΩ
÷5
SG
÷2
X1
X2
X5
R809
24kΩ
R826
10kΩ
R825
10kΩ
Channel B
zero
R808
24kΩ
−12V
Channel B
calibration
SF1
SF2
140 70 35
ms
C803
10µF
C804
10µF
Channel A
Channel B
Time/cm multiplier
Integrator output
(From TR602, Fig 17)
R816
500kΩ
R817
1MΩ
R818
2MΩ
R819
2MΩ
÷4
SH
Start at 0V
Time
÷2
X1
X2
R820
24kΩ
R827
10kΩ
Time
calibration
A ↑ T →
B ↑ T →
A ↑ B →
B ↑ A →
T ↑ A →
T ↑ B →
SI1
SI2
Y − Input on
plotter (↑)
25mV/cm,
27kΩ
X − Input on
plotter (→)
25mV/cm,
27kΩ
Display
mode
✱ Contacts 5 and 6 of RLC3 and RLC4
are closed until scan begins
Notes: (1) Oscilloscope outputs
are nominally 1V/cm
-of-c.r.t.-display,
and are via 10kΩ ± 10%
(2) All resistors are high-
stability types, ± 1%
or better
a3187

Fig. 17—Scale multipliers, filters and display-mode switching

Overload Sensing

Fig. 18 shows the overload detectors. Channel A and B detectors are fed from the differentiator emitter-follower and operate relays RLE and RLF respectively when the input voltages are <0V or >+12V, these levels corresponding to approximately ±9cm, beyond which overload of the sampling channels in the 3S76 occurs.

X- and Y-overload sensing is by voltages derived from the plotting table feedback potentiometers, and the modification to obtain these voltages is described in Appendix 1. Extreme left-hand or lower movement corresponds to nearly zero volts, whilst extreme right-hand or upper movement corresponds to −85V. The overload detectors of X- and Y-position are also shown in Fig. 19. The exact trip point is adjustable by R_{213} (left-hand), R_{214} (right-hand), R_{215} (lower) and R_{216} (upper). It is important not to load the feedback potentiometers whilst the carriage is inside the 'non-overload' area, and this is ensured with the circuit shown, by the high reverse-resistance of the diodes. Trip current is about 10µA.

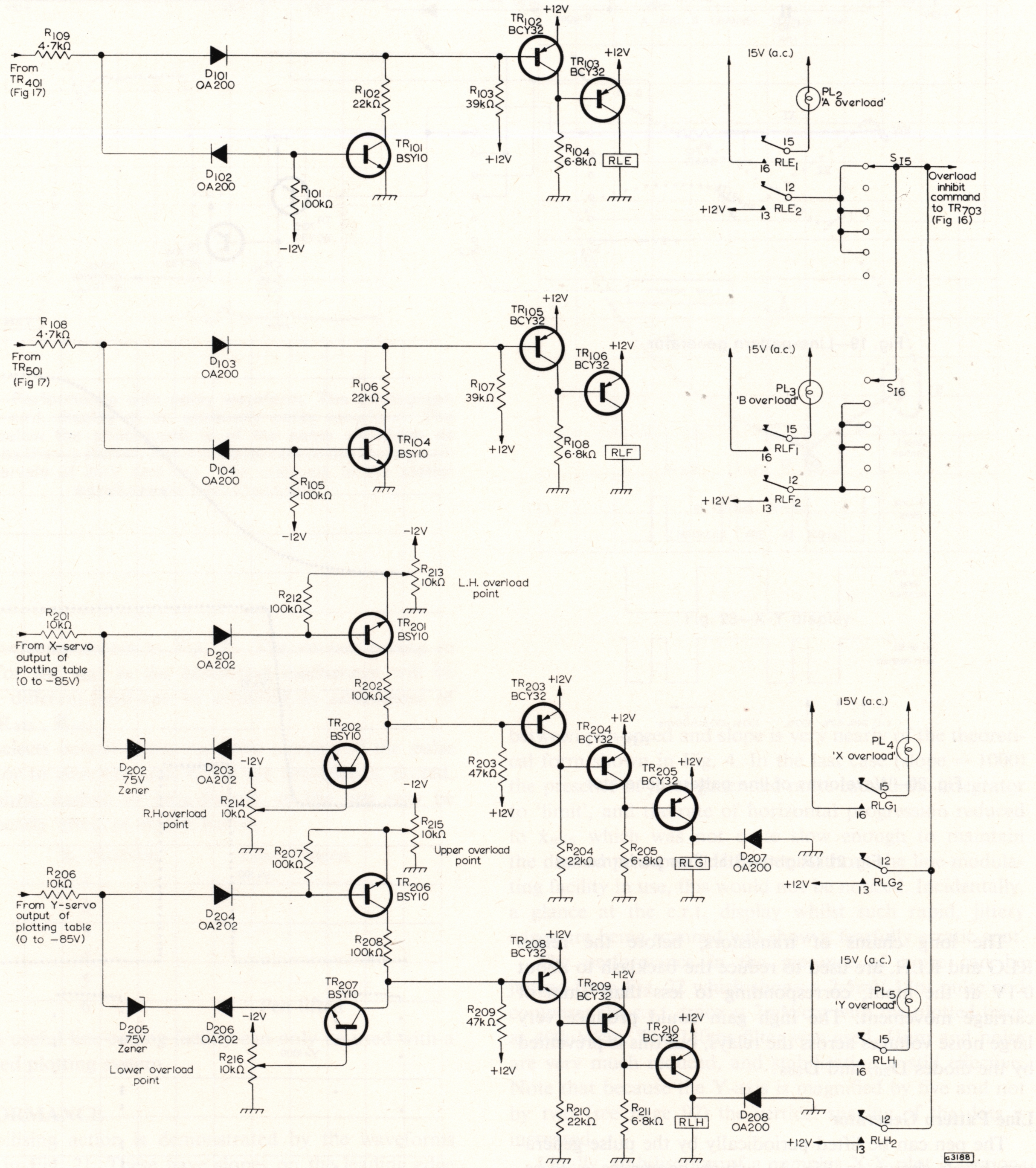

R109
4·7kΩ
From
TR401
(Fig 17)
D101
OA200
D102
OA200
R102
22kΩ
R101
100kΩ
−12V
TR101
BSY10
R103
39kΩ
+12V
TR102
BCY32
+12V
TR103
BCY32
+12V
R104
6·8kΩ
RLE
15V (a.c.)
PL2
'A overload'
15
RLE1
16
12
+12V
RLE2
13
SI5
Overload
inhibit
command
to TR703
(Fig 16)
R108
4·7kΩ
From
TR501
(Fig 17)
D103
OA200
D104
OA200
R106
22kΩ
R105
100kΩ
−12V
TR104
BSY10
R107
39kΩ
+12V
TR105
BCY32
+12V
TR106
BCY32
+12V
R108
6·8kΩ
RLF
15V (a.c.)
PL3
'B overload'
SI6
15
RLF1
16
12
+12V
RLF2
13
−12V
R213
10kΩ
L.H. overload
point
R212
100kΩ
R201
10kΩ
From X-servo
output of
plotting table
(0 to −85V)
D201
OA202
TR201
BSY10
R202
100kΩ
TR202
BSY10
D202
75V
Zener
D203
OA202
−12V
R214
10kΩ
R.H.overload
point
TR203
BCY32
+12V
R203
47kΩ
+12V
TR204
BCY32
+12V
TR205
BCY32
+12V
R204
22kΩ
R205
6·8kΩ
RLG
D207
OA200
15V (a.c.)
PL4
'X overload'
15
RLG1
16
12
+12V
RLG2
13
−12V
R215
10kΩ
Upper overload
point
R207
100kΩ
R206
10kΩ
From Y-servo
output of
plotting table
(0 to −85V)
D204
OA202
TR206
BSY10
R208
100kΩ
TR207
BSY10
D205
75V
Zener
D206
OA202
−12V
R216
10kΩ
Lower overload
point
TR208
BCY32
+12V
R209
47kΩ
+12V
TR209
BCY32
+12V
TR210
BCY32
+12V
R210
22kΩ
R211
6·8kΩ
RLH
D208
OA200
15V (a.c.)
PL5
'Y overload'
15
RLH1
16
12
+12V
RLH2
13
a3188

Fig. 18—Overload detectors

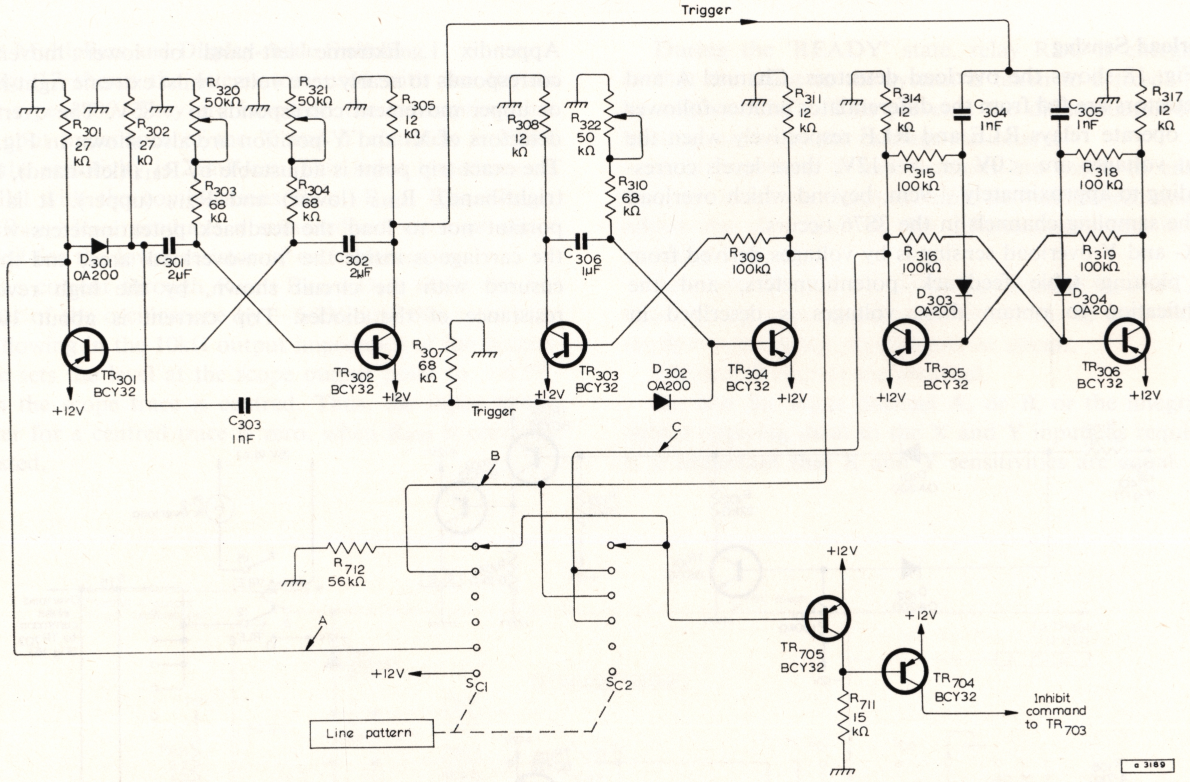

Trigger
R301 27 kΩ
R302 27 kΩ
R320 50kΩ
R321 50kΩ
R305 12 kΩ
R303 68 kΩ
R304 68 kΩ
D301 OA200
C301 2µF
C302 2µF
TR301 BCY32
TR302 BCY32
+12V
C303 1nF
R307 68 kΩ
Trigger
R308 12 kΩ
R322 50 kΩ
R310 68 kΩ
C306 1µF
R309 100kΩ
TR303 BCY32
D302 OA200
TR304 BCY32
R311 12 kΩ
R314 12 kΩ
C304 1nF
C305 1nF
R317 12 kΩ
R315 100 kΩ
R318 100 kΩ
R316 100kΩ
R319 100kΩ
D303 OA200
D304 OA200
TR305 BCY32
TR306 BCY32
B
C
A
R712 56kΩ
+12V
SC1
SC2
Line pattern
TR705 BCY32
R711 15 kΩ
TR704 BCY32
Inhibit command to TR703
a 3189

Fig. 19—Line-pattern generator

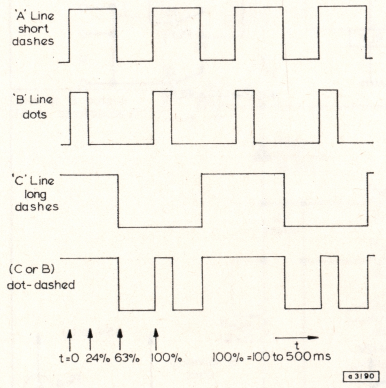

'A' Line short dashes
'B' Line dots
'C' Line long dashes
(C or B) dot-dashed
t=0 24% 63% 100%
100%+100 to 500ms
t
a3190

Fig. 20—Waveforms of line pattern generator

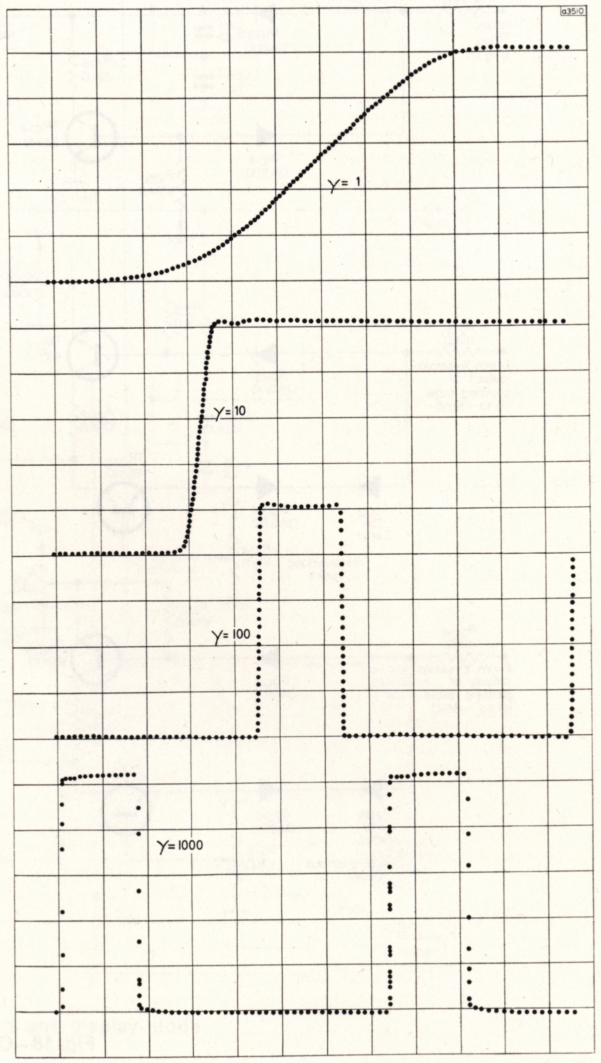


Fig. 21 (Right)—Stabiliser performance

The long chains of transistors, before the relays RLG and RLH, are used to reduce the backlash to about 0·1V at the input, corresponding to less than 1mm of carriage movement. The high gain could produce very large noise voltages across the relays, but this is prevented by the diodes D_{207} and D_{208}.

Line Pattern Generator

The pen can be lifted periodically by the pulse generating block shown in Fig. 19, which comprises a cross-coupled astable circuit, driving a cross-coupled bistable circuit and a cross-coupled monostable circuit to produce the waveforms shown in Fig. 20. The correct timing to allow for the lags in the pen-lifting mechanism will be slightly different, and can be achieved by adjustment of R_{320}, R_{321}, R_{322}.

S_C selects outputs from the three sections of the pulse generator to generate lines which are continuous, dotted, dot-dashed, dashed or long-dashed, or the pen may be permanently lifted, as shown below:

S_C *Position*	*Line Pattern*
1	――――――――
2	....
3	-.-.-.-.-.-.-.-
4	- - - - - - - - -
5	— — — —
6	Pen lifted

This useful line-coding facility can only be used with a stabilised plotting system.

PERFORMANCE

Stabilising action is demonstrated by the waveforms shown in Fig. 21. These have slopes on the leading edge of 1, 10, 100 and 1000, and the dots are spaced at approximately 200ms intervals. It can be seen that the relationship between pen-speed and slope is very nearly of the theoretical form shown in Fig. 4. In the last case (slope = 1000) the presence of jitter-generated noise caused the integrator to 'limit', and the rate of horizontal progression reduced to $\dot{x}_{\min}$ which was not quite slow enough to maintain the desired pen-speed. However, without the line-modulating facility in use, this would not be noticed. Incidentally, a glance at the c.r.t. display whilst such rapid, jittery edges are being scanned will show a fearfully erratic spot.

The performance in the presence of noise can be judged from Fig. 22 which shows a 2·5mV 18ns pulse and compares the c.r.t. trace at 2mV/cm with the plotted trace at 400µV/cm. The 140ms filter was used. Hum and noise are very much reduced, and stabilisation is still effective. Note that because the Y-axis is magnified by five and not by two (see page 92) the vertical spacing of the dots is increased.

Finally, stabilisation of a complex X-Y plot is demonstrated by Fig. 23. Although some regions (for example, III and IX) are plotted very slowly, the plotting speed

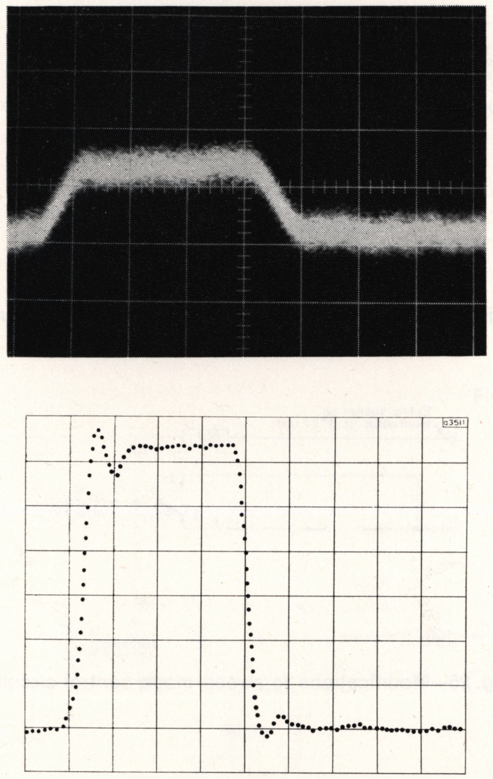


Fig. 22—Performance with noisy waveform. The photograph is of a c.r.t. display of an extremely noisy waveform. The curve below the photograph is of the same waveform as plotted by the x-y plotter. The main graticule in the photograph has intervals of 2mV and 5ns. The intervals in the plotter waveform are 0·4mV and 5ns.

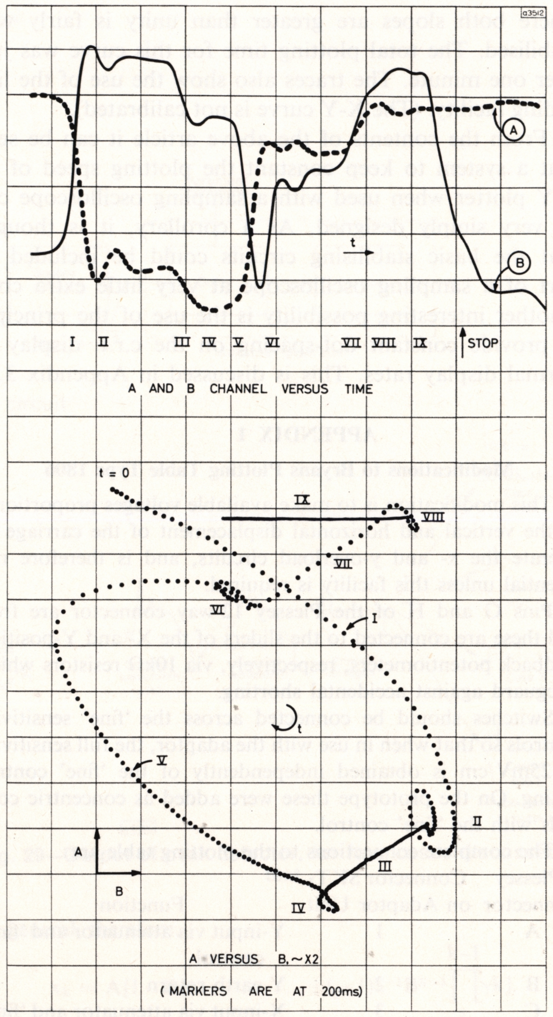


Fig. 23—X-Y display

where both slopes are greater than unity is fairly well stabilised. The total plotting time for this curve was just over one minute. The traces also show the use of the line coding facility. (The X-Y curve is not calibrated).

From the contents of the above article it can be seen that a system to keep constant the plotting speed of an X-Y plotter when used with a sampling oscilloscope can be very simply designed. As a corollary, it is thought that the basic stabilising circuits could be included as part of a sampling oscilloscope at very little extra cost. Another interesting possibility is the use of the principle to provide constant dot-spacing on the c.r.t. display at normal display rates. This is discussed in Appendix 3.

APPENDIX 1

Modifications to Bryans Plotting Table Type 1806

This modification is to make available voltages proportional to the vertical and horizontal displacement of the carriage to operate the x- and y-overload circuits, and is therefore not essential unless this facility is required.

Pins G and H of the Plessey 12-way connector are free, and these are connected to the sliders of the X- and Y-position feedback potentiometers, respectively, via 10kΩ resistors which safeguard against accidental shorting.

Switches should be connected across the 'fine' sensitivity controls so that when in use with the adaptor, the full sensitivity of 25mV/cm is obtained independently of the 'fine' control setting. On the prototype these were added as concentric controls with the 'fine' control.

The complete connections to the plotting table are:

Plessey Connector	Connector SKT_1 on Adaptor Unit	Function
A	1	Y-input via attenuator and 'fine' control
B	2	Y earth return
C	3	X-input via attenuator and 'fine' control
D	4	X earth return
E	5	Y-input direct, 27kΩ, 25mV/cm
F	6	X-input direct, 27kΩ, 25mV/cm
G	7	X position voltage, 0 to −85V
H	8	Y position voltage, 0 to −85V
J	9	short to fire timebase on plotting table
K	10	
L	11	short to lower pen (pin 11 = zero, pin 12 = +37V; 'on' current = 23mA)
M	12	

APPENDIX 2

Modification to Tektronix 3T77 unit

This simple modification is essential if automatic transfer from the direct display to the plotting mode is required. The original circuit around the 'SWEEP MODE' control is shown in Fig. 24, which is on the left-hand side of the Horizontal Amplifier schematic in the 3T77 manual. When SW145 is switched to 'EXTERNAL SWEEP INPUT', the control R_{301} determines what fraction of the input voltage is applied to the rest of the system. When driven from the adaptor it is important that this control cannot interfere with the scanning voltage applied to the rest of the system. The circuit of Fig. 25, incorporating the relay contact RLJ_1 on RLJ, meets the system's requirements. Whichever position SW_{145} is in before the 'START' button on the adaptor is depressed, this control is over-ridden.

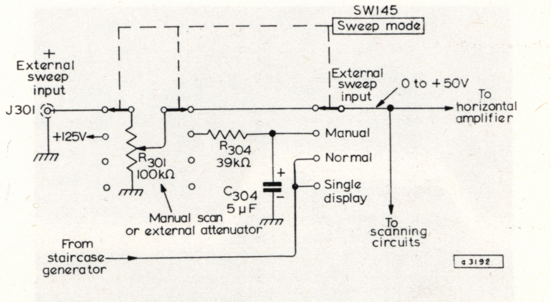

SW145
Sweep mode
External sweep input
J301
+125V
R301 100kΩ
R304 39kΩ
C304 5μF
Manual scan or external attenuator
External sweep input
0 to +50V
To horizontal amplifier
Manual
Normal
Single display
To scanning circuits
From staircase-generator
a3192

Fig. 24—Original circuit of sweep-mode control circuit

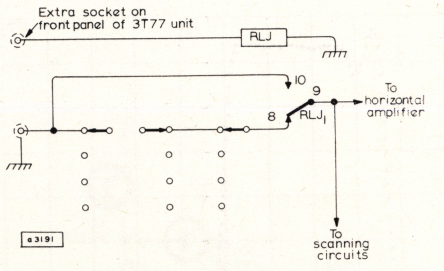

Extra socket on front panel of 3T77 unit
RLJ
10
9
8
RLJ1
To horizontal amplifier
To scanning circuits
a3191

Fig. 25—Modifications to sweep-mode control circuit

APPENDIX 3

A Discussion of the Possibilities for Stabilising the Internal Scanning System of a Sampling Oscilloscope

Consider Fig. 26, which shows a hypothetical waveform in the X, Y(X) plane, or the t, Y(t) plane, sampled at points, 1, 2, . . . K, K+1.

x_k is the space (or time) increment between sample K and K+1, and y_k is the corresponding space (or voltage) increment. In practice, the integral distance X is the output from the staircase generator, whilst Y is the output from the vertical memory and y_k is the output from the vertical error amplifier.

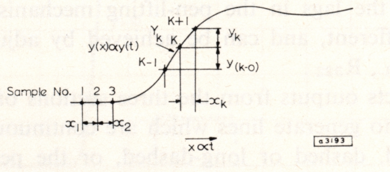

K+1
Y K
K
y(x)∝y(t)
K-1
y k
y (k-o)
Sample No. 1 2 3
x1
x2
x k
x∝t
a3193

Fig. 26—Hypothetical waveform

Notice that the increment x_k is added to X when sample K is made, but y_k does not appear until sample K+1.

In the proposed system, we must control x_k by $y_{(k-1)}$ such that

$$x_{(k-1)}^2+y_{(k-1)}^2 = x_k^2+y_k^2 = \text{const} = A, \qquad \ldots(6)$$

that is, arrange for the arc-length to be constant.

This gives

$$x_k = \sqrt{(A-y_k^2)}, \qquad \ldots(7)$$

which obviously cannot be realised since y_k is not known until sample K+1 has been executed. However, if we assume that Y_K, the waveform slope at sample K, is varying only slowly, then

$$x_k = \sqrt{(A-y_{(k-1)}^2)} \qquad \ldots(8)$$

is a good approximation. This function can be realised by the

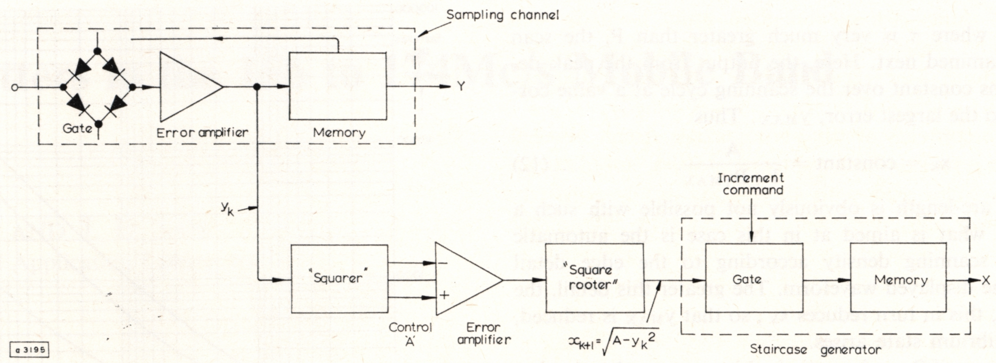

Sampling channel
Gate
Error amplifier
Memory
Y
Y k
Increment command
"Squarer"
Control 'A'
Error amplifier
"Square rooter"
$x_{k+1} = \sqrt{A-y_k^2}$
Gate
Memory
X
Staircase generator
a3195

Fig. 27—Required circuit

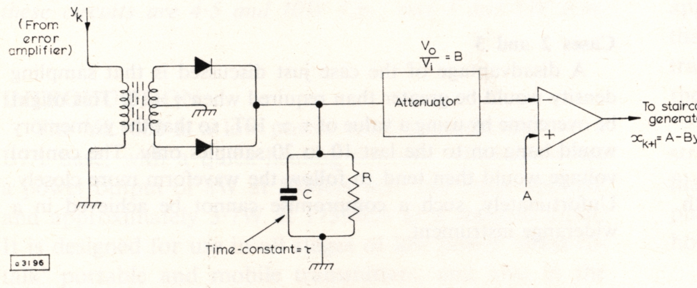

(From error amplifier)
Y k
$\frac{V_o}{V_i}=B$
Attenuator
To staircase generator
$x_{k+1}=A-By_k e^{-t/\tau}$
A
C
R
Time-constant=τ
a3196

Fig. 28—Alternative circuit

circuit shown in Fig. 27, but is likely to be too expensive for general use, unless the square-law functions could be performed using silicon-carbide or similar non-linear resistance elements.

Another possibility is to use a system such as the one shown in Fig. 28, which consists of a peak-detector circuit measuring y_k followed by a memory of time-constant τ which controls the staircase generator, according to the law:

$$x_k = A - By_{(k-1)}e^{-t/\tau} \qquad \ldots(9)$$

How the system behaves depends on the value of τ and the sampling interval T, and, if τ is very large, on the scan period, P. The peak-detector outputs are sketched in Fig. 29 for the four cases:

(i) $\tau \ll T$
(ii) $\tau \simeq T$
(iii) $\tau > T$
(iv) $\tau \gg T$, or $\tau \gg P$.

Case 1

In this case, y_{k-2} decays completely before the next sample K-1, is made. Thus x_k is based entirely on the value of y_{k-1}:

$$x_k = A - By_{(k-1)}. \qquad \ldots(10)$$

Consider the sequence of events at the start of cycle, assuming $y_{(k<1)} = 0$.

$$x_1 = A$$
$$x_2 = A - By$$

but

$$y_1 = x_1\gamma_1.$$

Therefore

$$\begin{aligned} x_2 &= A(1-B\gamma_1)\\ x_3 &= A-By_2\\ &= A(1-B\gamma_2-B^2\gamma_1\gamma_2). \end{aligned}$$

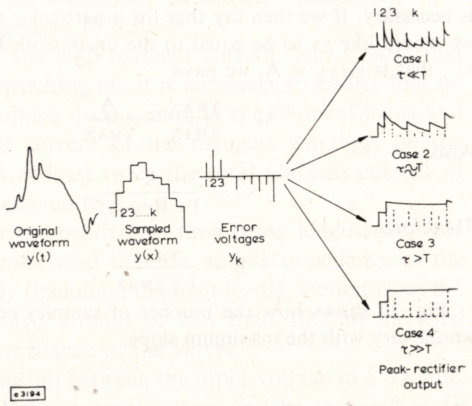

1 2 3 k
Case 1
τ≪T
Case 2
τ≃T
Case 3
τ>T
Case 4
τ≫T
Peak-rectifier output
Original waveform y(t)
1 2 3 ... k
Sampled waveform y(x)
1 2 3
Error voltages Y k
a3194

Fig. 29—Output of peak detector, for four cases discussed in text

In general, therefore,

$$x_k = A[1-B\gamma_{k-1}+\ldots\ldots+(-1)^{k-1}B^{k-1}\prod_{1}^{k-1}\gamma k]. \qquad \ldots(11)$$

There are two aspects of stabilisation. These are

(i) $x_k^2+y_k^2 = \text{const}$, for any value of γ_k.

(ii) $x_k = x_{k+1}$, when γ is constant between samples K and K+2.

The former ensures constant dot-density, as previously mentioned, whilst the latter indicates that stabilisation has reached an equilibrium state even though the first condition is not satisfied.

For the latter condition to be true

$$(-1)^{k-1}B^{k-1}\prod_{1}^{k-1}\gamma_k = (-B\gamma)^{k-1} = 0.$$

Thus, for convergence to occur at all, $B\gamma<1$, and for it to occur within a small number of samples, $B\gamma \ll 1$. Theoretically, the system must stabilise for infinite γ, so that B must be zero; this being so, x_k is invariant and no stabilisation is possible.

Also, even allowing an infinite time for stabilisation to occur (that is $B\gamma = 1$), the maximum stable reduction in x_k is from A to A/2. This can be easily seen, since,

$$x_k = \frac{A}{1+B\gamma_k} \text{ (from Eq(10))}.$$

Case 4

The case where τ is very much greater than P, the scan period, is examined next. Here, the output from the peak detector remains constant over the scanning cycle at a value corresponding to the largest error, y_{MAX}. Thus

$$x_k = \text{constant} = \frac{A}{1+B\gamma_{MAX}}. \qquad \ldots(12)$$

Constant arc-length is obviously not possible with such a method, but what is aimed at in this case is the automatic selection of scanning density according to the edge detail present on the displayed waveform. The greater this detail, the larger y_{MAX} ; this in turn reduces x_k , so that y_{MAX} is reduced, and an equilibrium state arises.

Let us now see if B can be optimised in any way. A reduction in dot-density is not very important if the definition does not suffer too much. This usually means that for $\gamma<10$, no control is necessary. If we then say that for a particular value of γ_{MAX} we would like y_k to be equal to the uncontrolled increment in x_k , that is $y_{MAX} = A$, we have

$$x_k = \frac{y_{MAX}}{\gamma_{MAX}} = \frac{A}{\gamma_{MAX}}.$$

Also,

$$x_k = \frac{A}{1+B\gamma_{MAX}}.$$

Therefore

$$B = \frac{\gamma_{MAX}}{1+\gamma_{MAX}}.$$

Fig. 30 shows how the number of samples per unit length would vary with the maximum slope.

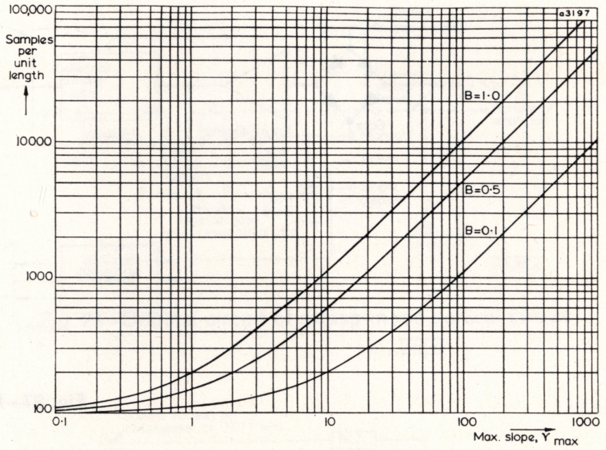


Fig. 30—Variation of number of samples per unit length with value of maximum slope

Cases 2 and 3

A disadvantage of the case just discussed is that sampling density would be greater than required when $\gamma = 0$. This might be overcome by using a value of $\tau \simeq 10T$, so that the y_k memory would hang on to the last 10 to 20 samples only. The control voltage would then tend to follow the waveform more closely. Unfortunately, such a compromise cannot be achieved in a widerange instrument.

Transmitters in the 156 to 174Mc/s Mobile Band

L. M. Cash and J. E. Saw
Mullard Central Application Laboratory

The YL1000, from the comprehensive range of Mullard quick-heating valves, provides 70% output power within 0·5 seconds of switching on. A general introduction to the applications of the YL1000 is given, together with four circuits for mobile transmitters. The outputs provided by these circuits are 4·5 and 10W f.m., and 3 and 6W a.m.

INTRODUCTION

The YL1000 is a low power v.h.f. pentode built on a B9A base, with a quick-heating filament which provides a power output of 8W at 50Mc/s with full input applied, and approximately 3·7W at 175Mc/s at a reduced input. It is designed for use in all stages of low power 'press-to-talk' portable and mobile transmitters, and also in the master oscillator, frequency multiplier, driver and power amplifier stages of high power equipment.

In a typical equipment (Fig. 4) containing five YL1000 valves, with a total filament supply circuit impedance of 0·02Ω, 70% of the power output is obtained in 0·5 seconds or within 1 second if the filament voltage is 15% below nominal.

Where only the YL1000 is used throughout a transmitter, or if it is used in conjunction with other quick-heating valves, the filament voltage may be switched off during listen-out periods.

This facility, which is of considerable importance in portable and mobile equipment where semiconductor devices are used in the receiver and power supply, provides a substantial economy in the average power consumption.

FILAMENT SUPPLY VOLTAGE

The filament of the YL1000 operates at a low voltage and high current (1·1V, 0·88A) and cannot be operated directly from a 6 or 12V car battery.

The connection of the filaments of several valves in series with a dropping resistor is not practicable, since the heating-up time would be considerably lengthened and power would be lost in the resistor.

The most suitable method of obtaining the a.c. filament voltage is from additional windings on the transformer of a transistor d.c./d.c. converter. The output from these windings is a rectangular shaped, alternating voltage, the r.m.s. value of which is half the peak-to-peak amplitude.

Because of the shape of the waveform, the filament voltage should be measured by means of an oscilloscope or moving iron voltmeter, since the published figure refers to the r.m.s. or d.c. value.

FILAMENT VOLTAGE TOLERANCES

The absolute maximum permitted variation of the filament voltage is ±15%.

Because of the high filament current, particularly that drawn when switching on, it is necessary to ensure that no appreciable voltage drop occurs in the filament leads and that the rated current of the filament windings on the transformer is at least twice the total filament current of the valves connected to it.

To prevent the heating-up time being lengthened, it is strongly recommended that the source impedance of the filament supply (including filament leads), viewed from the pins of any valve, should be less than 10% of the total hot filament resistance of the valves.

The relationship between the input voltage to a d.c./d.c. converter and the output voltage can be assumed to be linear, providing the impedance of the input source is low. The filament voltage of the valves therefore changes linearly with the battery voltage.

METHODS OF FILAMENT CONNECTION

The filament of a directly-heated valve is not at a uniform potential but has a potential gradient along its length. When the filament is fed with an alternating voltage the potential gradient along the filament will alternate. If the electrode circuit connections are returned to one side of the filament, the voltage on each electrode with respect to the filament will fluctuate at the frequency of the filament supply voltage. These 'hum' voltages have the same effect as an equivalent external voltage applied between an electrode and the filament and will modulate the anode current. The most significant hum voltage is that between the control grid and filament.

To minimise the effect of hum voltages, a voltage divider should be connected across the filament supply. The electrical centre of the voltage divider is used as the return point for the electrode circuit connections. This is equivalent to connecting them to the effective centre or equipotential point of the filament, because the alternating voltages between the centre point and each side of the filament are now equal and opposite.

If an equipment is to comply with the published U.K. regulations concerning spurious signals (H.M.S.O. ref. S.O.43-179), considerable attention must be paid to the design arrangement of the filament circuit and the choice of a.c. filament supply frequency.

DESIGN OF THE FILAMENT CIRCUIT

The filament circuit may be arranged in one of three ways: with the centre-tap supply to earth, with a humdinging arrangement, or with one side of the filament earthed. Measurements have been made on each of these arrangements to determine the total hum modulation products of the YL1000, operating as a power amplifier at 28Mc/s with a 200c/s sine wave filament supply. The results obtained are relative to the carrier amplitude and are as follows:

with a suitable humdinging arrangement	−70dB
with a centre-tapped filament winding for a transformer made to a commercial specification	−50dB
with one side of the filament earthed	−40dB

These results were obtained with accurate tuning, and optimum drive voltage and coupling between stages.

Slight detuning, with a consequent reduction in drive voltage—especially in the grid circuit of an output stage—can increase the hum modulation by approximately 10dB. No significant difference was observed when the filament waveform was changed to a square wave.

Where several stages employing humdinging are used, better figures can be achieved, since it is possible to cancel some of the residual hum in one stage by adjustment of the humdinging arrangement on a previous stage.

Care should be taken in positioning the converter and the supply battery relative to the transmitter chassis and the connecting leads, to ensure that no induced voltages and consequent hum modulation arise from these sources.

A typical humdinging arrangement is shown in Fig. 1.

CHOICE OF FILAMENT SUPPLY FREQUENCY

The frequency of the hum modulation voltage is either the fundamental or a harmonic of the filament supply frequency. The choice of whether a low or high frequency should be used for the filament supply is governed by several factors. Both have disadvantages and although no conclusive case can be made for either, a low frequency seems to be preferable.

The advantages and disadvantages of low and high frequencies are summarised below.

Low frequency

A larger core is required for the transformer of the d.c./d.c. converter.

More expensive filament filtering components are required.

It is difficult to choose a frequency at which the fundamental, or any harmonic, does not coincide with the filament resonance frequencies of the various types of valves in an equipment.

Providing the filtering is sufficient, the hum voltage only produces modulation products which are audible and no adjacent channel interference occurs.

High frequency

A cheaper transformer core and filter components can be used.

No audible hum modulation is produced but harmonics cause adjacent channel interference.

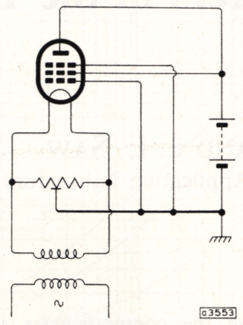


Fig. 1—Typical humdinging arrangement

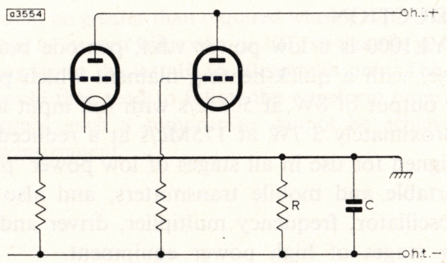


Fig. 2—Principle of partial fixed bias system

The frequency of the filament supply can be chosen to be higher than the filament resonant frequency of the valves.

PROTECTION CIRCUIT

In a transmitter it is necessary to protect the valves from damage caused by loss of grid drive voltage. This may be done by providing a proportion of the bias voltage for each valve from a fixed supply. The circuit shown in Fig. 2 illustrates a suitable arrangement. A resistor (R) is inserted between the chassis earth line and the negative h.t. line. By returning the control grid resistors of each of the valves to the negative h.t. line the voltage developed across R forms part of the total bias voltage of each valve.

The voltage across R should be sufficient to prevent the maximum ratings of any valve being exceeded should the grid drive voltage be removed. In selecting the valve operating conditions allowance must be made for the voltage drop across R, and if necessary the h.t. supply voltage should be increased by this amount.

In the circuits shown in Figs. 3 to 6, the value of the control grid resistors should be modified if a protection circuit is used. For example, in the 3W transmitter for a.m. shown on page 109, R should be 330Ω, 3W and the control grid resistors of V2, V3 and V4 should be 270kΩ, 150kΩ and 82kΩ respectively. The h.t. voltage should be increased from 200 to 220V to allow for the 20V dropped across R.

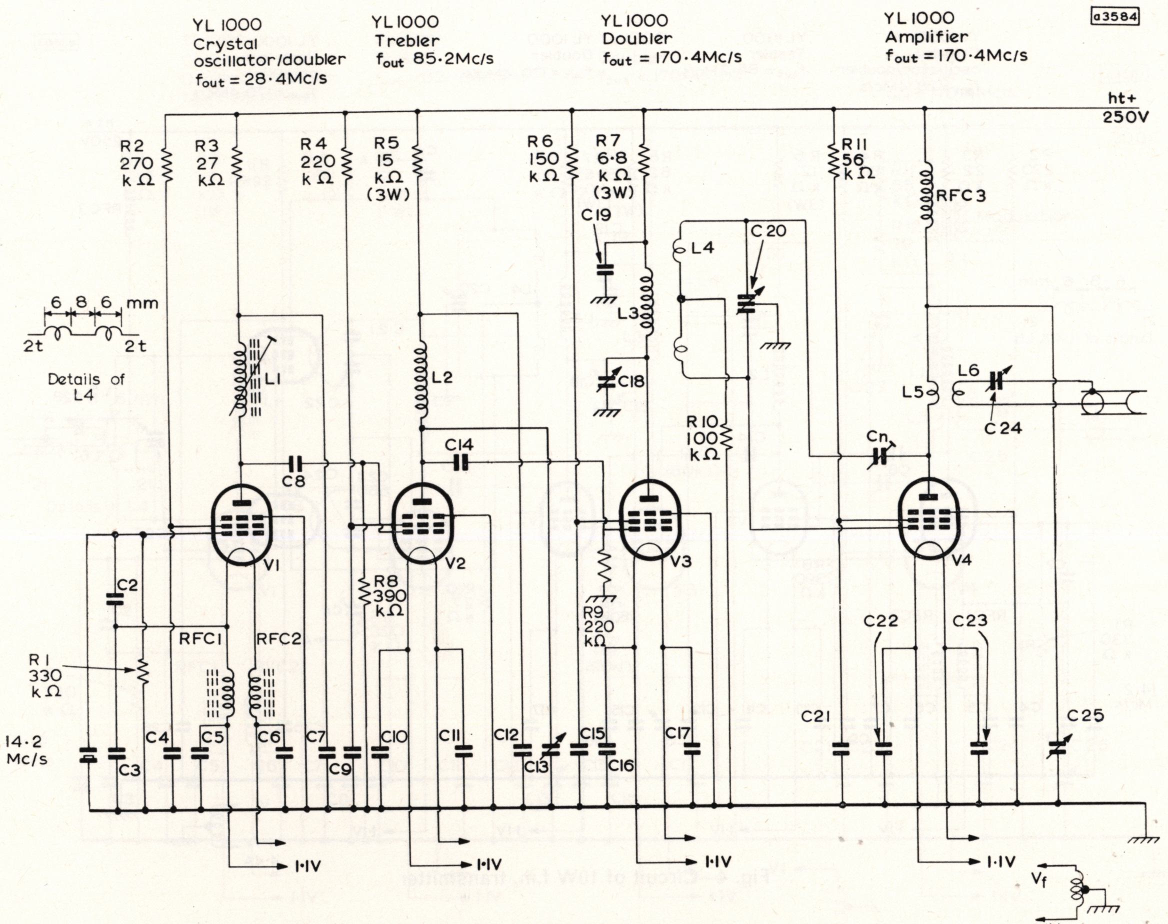


Fig. 3—Circuit of 4·5W f.m. transmitter

4·5W TRANSMITTER FOR F.M.

Stage 1

This is a crystal-controlled electron coupled oscillator-doubler. The filament, and the control and screen grids operate as a grounded-anode Colpitts oscillator at the crystal frequency of 14·2Mc/s. The anode circuit is tuned to 28·4Mc/s, which is twice the crystal frequency. (See Fig. 3.)

Stage 2

This stage, which is capacitively coupled to the oscillator, operates as a frequency trebler. The output circuit is tuned to 85·2Mc/s.

Stage 3

This stage, capacitively coupled to the trebler, operates as a frequency doubler. The output circuit is tuned to 170·4Mc/s and is inductively coupled to the input circuit of the final stage.

Stage 4

The fourth and final stage is a neutralised amplifier. The input circuit uses a split-stator capacitor to facilitate neutralising. The output circuit is series-tuned. This enables a larger coil to be used in the output circuit and also gives approximately 11% more load power than the conventional parallel-tuned circuit.

Approximate valve operating conditions

	V_f (V)	V_b (V)	V_a (V)	V_{g2} (V)	$-V_{g1}$ (V)	I_a (mA)	I_{g2} (mA)	I_{g1} (µA)
V_1	1·1	250	159	64	26	3·4	0·69	79
V_2	1·1	250	108	90	70	9·5	0·73	180
V_3	1·1	250	131	120	80	17·5	0·87	360
V_4	1·1	250	250	140	40	36	1·96	400

Coils

	No. of turns	Wire diam. (s.w.g.)	Material	Internal coil diam. (mm)	Length (mm)
L_1	24	26	Enamelled copper	7	close-wound (slug tuned)
L_2	5	14	Copper	11·5	22
L_3	1	14	Copper	11·5	—
L_4	2+2	14	Copper	11·5	20 (see Fig. 3)
L_5	5	15	Copper	13	20
L_6	1	14	Copper	13	—
RFC_1 RFC_2	15	26	Enamelled copper	Wound on ferrite core 12×3mm	
RFC_3	Painton miniature choke 200/220				

Resistors

All resistors ±20%, ½W except the following which are ±10%
R_2, R_3, R_4, R_5 (3W), R_6, R_7 (3W), R_{11}

Capacitors (all values in pF)

C_2	8·2	C_{11}	1000	C_{20}	4 to 18 per section
C_3	120	C_{12}	1000		
C_4	1000	C_{13}	3·5 to 21	C_{21}	1000
C_5	1000	C_{14}	47	C_{22}	1000
C_6	1000	C_{15}	1000	C_{23}	1000
C_7	2700	C_{16}	1000	C_{24}	3 to 30
C_8	10	C_{17}	1000	C_{25}	3·5 to 21
C_9	1000	C_{18}	3·5 to 21	C_n	Neutralising wire
C_{10}	1000	C_{19}	1000		

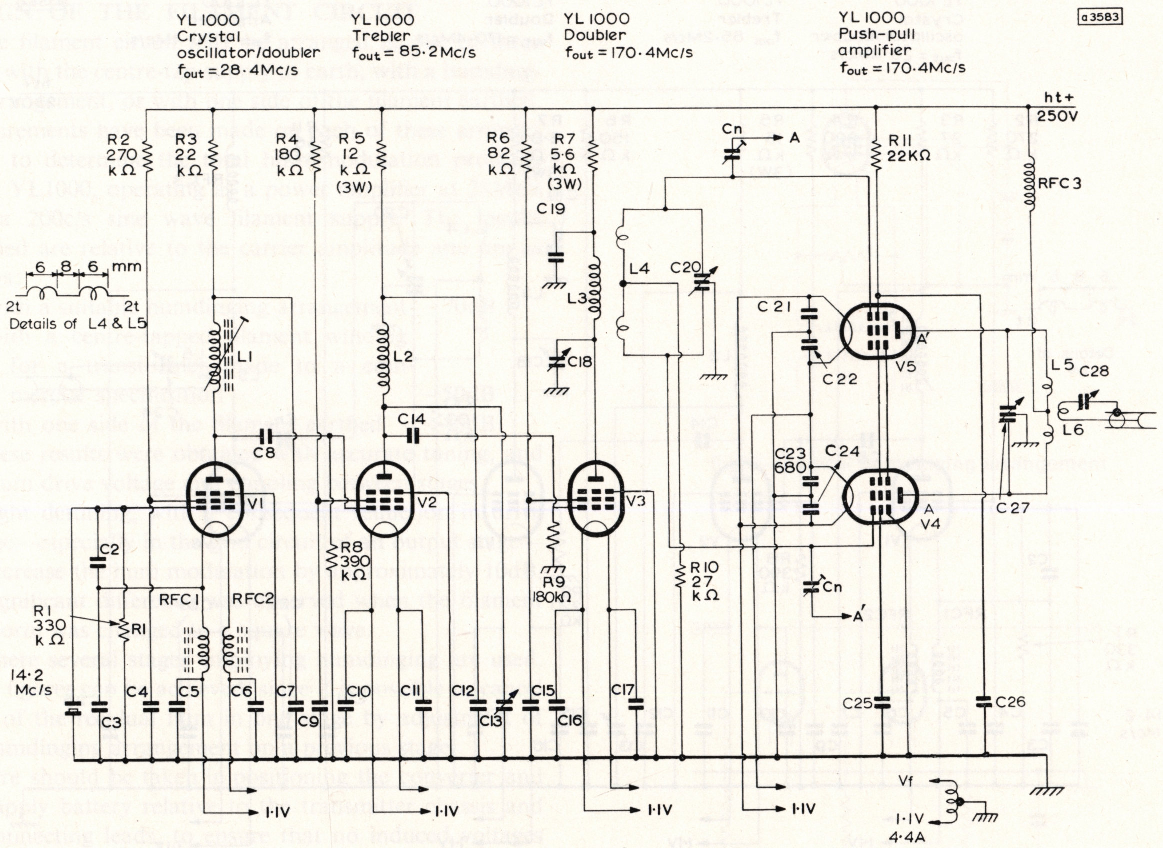

a3583
YL1000 Crystal oscillator/doubler fout = 28·4Mc/s
YL1000 Trebler fout = 85·2Mc/s
YL1000 Doubler fout = 170·4Mc/s
YL1000 Push-pull amplifier fout = 170·4Mc/s
ht+ 250V
R2 270 kΩ
R3 22 kΩ
R4 180 kΩ
R5 12 kΩ (3W)
R6 82 kΩ
R7 5·6 kΩ (3W)
R11 22kΩ
RFC3
6 8 6 mm
2t 2t
Details of L4 & L5
L1
L2
L3
L4
C18
C19
C20
Cn
A
C21
C22
V5
A'
L5
C28
L6
C14
C8
V1
V2
V3
C23 680
C24
V4
A
C27
C2
R8 390 kΩ
R9 180kΩ
R10 27 kΩ
Cn
A'
R1 330 kΩ
RFC1
RFC2
14·2 Mc/s
C3
C4
C5
C6
C7
C9
C10
C11
C12
C13
C15
C16
C17
C25
C26
Vf
1·IV
4·4A

Fig. 4—Circuit of 10W f.m. transmitter

10W TRANSMITTER FOR F.M.

The first three stages V1, V2 and V3 are virtually identical to those in the 4·5W transmitter. The output stage consists of two cross-neutralised YL1000 operating as a push-pull amplifier. (See Fig. 4.)

Coils

	No. of turns	Wire diam. (s.w.g.)	Material	Internal coil diam. (mm)	Length (mm)
L_1	24	26	Enamelled copper	7	close-wound (slug tuned)
L_2	5	14	Copper	11·5	18
L_3	2	14	Copper	11·5	—
L_4 L_5	2+2	14	Copper	11·5	20 (see Fig. 4)
L_6	1½	14	Copper	11·5	—
RFC_1 RFC_2	15	26	Enamelled copper	Wound on ferrite core 12×3mm	
RFC_3	Painton miniature choke 200/220				

Resistors

All resistors are ±20%, ½W except the following which are ±10%.

R_2
R_3
R_4
R_5, 3W
R_6
R_7, 3W
R_{11}

Capacitors (all values in pF)

C_2	8·2	C_{13}	3 to 21	C_{23}	1000
C_3	120	C_{14}	56	C_{24}	1000
C_4	2700	C_{15}	1200	C_{25}	1000
C_5	1000	C_{16}	1000	C_{26}	1000
C_6	1000	C_{17}	1000	C_{27}	3 to 14 per section
C_7	2700	C_{18}	2 to 12		
C_8	12	C_{19}	1000	C_{28}	3 to 30
C_9	1000	C_{20}	3 to 14 per section	C_n	Neutralising wire
C_{10}	1000				
C_{11}	1000	C_{21}	1000		
C_{12}	1000	C_{22}	1000		

Approximate valve operating conditions

	V_f (V)	V_b (V)	V_a (V)	V_{g2} (V)	$-V_{g1}$ (V)	I_a (mA)	I_{g2} (mA)	I_{g1} (μA)
V_1	1·1	250	172	64	18	3·54	0·69	55
V_2	1·1	250	115	98	84	11·3	0·85	210
V_3	1·1	250	133	124	83	20·9	1·54	460
V_4 V_5	1·1	250	250	144	38	2×40	2×2·4	2×675

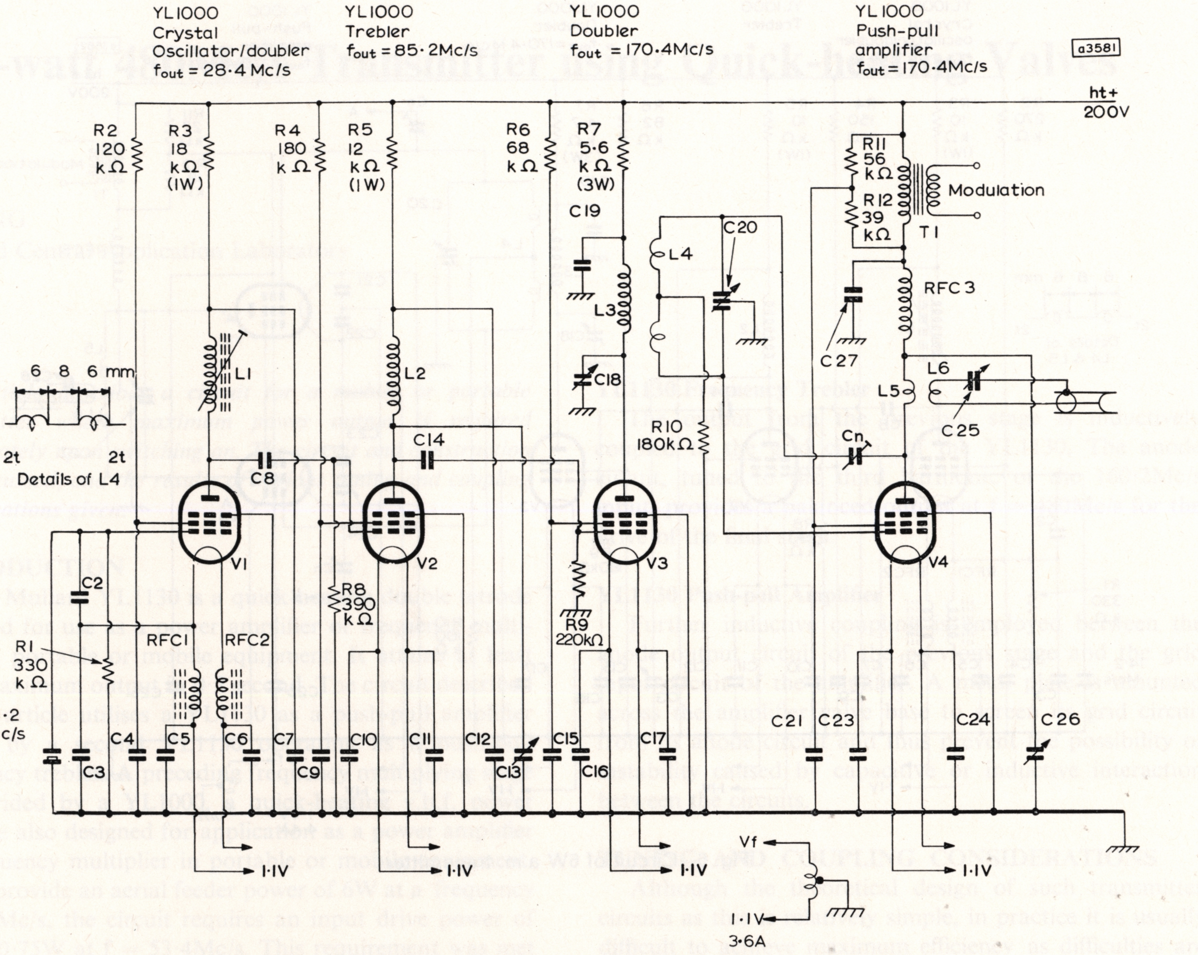

a3581
YL1000 Crystal Oscillator/doubler fout = 28·4Mc/s
YL1000 Trebler fout = 85·2Mc/s
YL1000 Doubler fout = 170·4Mc/s
YL1000 Push-pull amplifier fout = 170·4Mc/s
ht+ 200V
R2 120 kΩ
R3 18 kΩ (1W)
R4 180 kΩ
R5 12 kΩ (1W)
R6 68 kΩ
R7 5·6 kΩ (3W)
R11 56 kΩ
R12 39 kΩ
Modulation
T1
C19
C20
L4
L3
C18
RFC3
C27
L6
L5
C25
Cn
R10 180kΩ
6 8 6 mm
2t 2t
Details of L4
L1
L2
C14
C8
V1
V2
V3
V4
C2
R8 390 kΩ
R9 220kΩ
R1 330 kΩ
RFC1
RFC2
14·2 Mc/s
C3
C4
C5
C6
C7
C9
C10
C11
C12
C13
C15
C16
C17
C21
C23
C24
C26
Vf
1·IV
3·6A

Fig. 5—Circuit of 3W a.m. transmitter

3W TRANSMITTER FOR A.M.

This circuit is derived from the 4·5W f.m. circuit and has provision for anode and screen-grid modulation. (See Fig. 5.)

Coils

	No. of turns	Wire diam. (s.w.g.)	Material	Internal coil diam. (mm)	Length (mm)
L_1	24	26	Enamelled copper	7	close-wound (slug tuned)
L_2	5	14	Copper	11·5	22
L_3	2	14	Copper	11·5	—
L_4	2+2	14	Copper	11·5	20 (see Fig. 5)
L_5	5	14	Copper	13	20
L_6	1	14	Copper	13	—
RFC_1 RFC_2	15	26	Enamelled copper	Wound on ferrite core 12×3mm	
RFC_3	Painton miniature choke 200/220				

Resistors

All resistors are ±20%, ½W except the following which are ±10%.

R_2
R_3, 1W
R_4
R_5, 1W
R_6
R_7, 3W
R_{11}
R_{12}

Capacitors (all values in pF)

C_2	8·2	C_{12}	1000	C_{21}	1000
C_3	120	C_{13}	3·5 to 21	C_{23}	1000
C_4	2700	C_{14}	47	C_{24}	1000
C_5	1000	C_{15}	1000	C_{25}	3 to 30
C_6	1000	C_{16}	1000	C_{26}	3·5 to 21
C_7	2700	C_{17}	1000	C_{27}	1000
C_8	10	C_{18}	2 to 12	C_n	Neutralising wire
C_9	1000	C_{19}	1000		
C_{10}	1000	C_{20}	4 to 18 per section		
C_{11}	1000				

Approximate valve operating conditions

Carrier conditions for 100% modulation

	V_f (V)	V_b (V)	V_a (V)	V_{g2} (V)	$-V_{g1}$ (V)	I_a (mA)	I_{g2} (mA)	I_{g1} (μA)
V_1	1·1	200	137	93	12	3·5	0·77	37
V_2	1·1	200	96	85	62	8·7	0·77	160
V_3	1·1	200	117	99	70	14·8	1·8	330
V_4	1·1	200	200	150	40	32	2·2	200

$P_{mod} = 3·5W$ This is the power required at the secondary of the modulation transformer, and allowance should be made for the efficiency of this transformer.

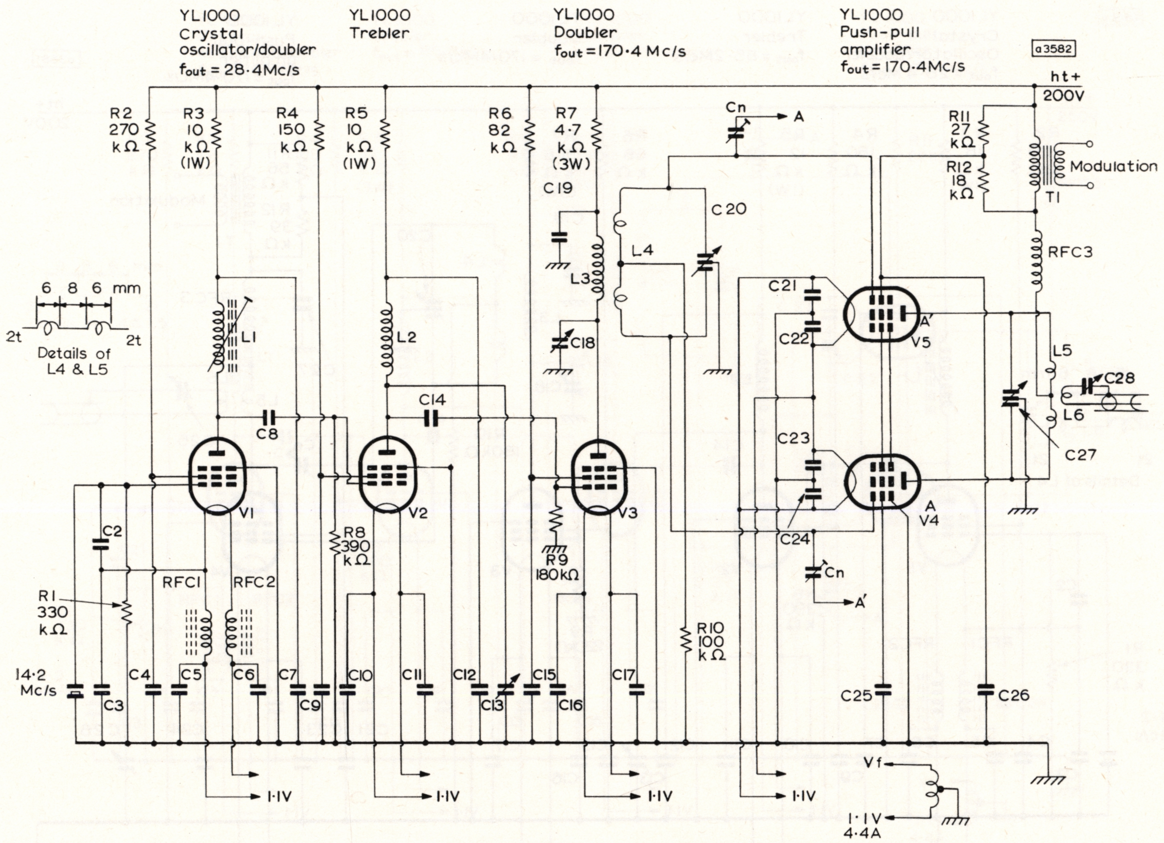


Fig. 6—Circuit of 6W a.m. transmitter

6W TRANSMITTER FOR A.M.

This circuit is derived from the 10W f.m. circuit and has provision for anode and screen-grid modulation. (See Fig. 6.)

Coils

	No. of turns	Wire diam. (s.w.g.)	Material	Internal coil diam. (mm)	Length (mm)
L_1	24	26	Enamelled copper	7	close-wound (slug tuned)
L_2	5	14	Copper	11·5	18
L_3	2	14	Copper	11·5	—
L_4 L_5	2+2	14	Copper	11·5	20 (see Fig. 6)
L_6	1½	14	Copper	11·5	—
RFC_1 RFC_2	15	26	Enamelled copper	Wound on ferrite core 12×3mm	
RFC_3	Painton miniature choke 200/220				

Resistors

All resistors are ±20%, ½W except following which are ±10%.

R_2	R_6
R_3 1W	R_7 3W
R_4	R_{11}
R_5 1W	R_{12}

Capacitors (all values in pF)

C_2	8·2	C_{13}	3 to 21	C_{23}	1000
C_3	120	C_{14}	56	C_{24}	1000
C_4	2700	C_{15}	1000	C_{25}	1000
C_5	1000	C_{16}	1000	C_{26}	1000
C_6	1000	C_{17}	1000	C_{27}	3 to 14 per section
C_7	2700	C_{18}	2 to 12		
C_8	12	C_{19}	1000	C_{28}	3 to 30
C_9	1000	C_{20}	3 to 14 per section	C_n	Neutralising wire
C_{10}	1000				
C_{11}	1000	C_{21}	1000		
C_{12}	1000	C_{22}	1000		

Approximate valve operating conditions

Carrier conditions for 100% modulation

	V_f (V)	V_b (V)	V_a (V)	V_{g2} (V)	$-V_{g1}$ (V)	I_a (mA)	I_{g2} (mA)	I_{g1} (µA)
V_1	1·1	200	164	78	12	3·6	0·68	36
V_2	1·1	200	117	92	77	8·3	0·72	200
V_3	1·1	200	116	104	82	18	1·17	460
V_4 V_5	1·1	200	200	154	40	2×31	2×2·3	400

P_{mod} = 6·9W This is the power required at the secondary of the modulation transformer, and allowance should be made for the efficiency of this transformer.

Six-watt 480Mc/s Transmitter using Quick-heating Valves

J. Ling
Mullard Central Application Laboratory

This article describes a circuit for a mobile or portable transmitter where maximum power output is required immediately upon switching on. The circuit and construction are discussed, and the results of various tuning and coupling investigations given.

INTRODUCTION

The Mullard YL1130 is a quick-heating double tetrode designed for use as a power amplifier or frequency multiplier in portable or mobile equipment. It attains at least 70% maximum output in 0·5 second. The circuit described in this article utilises a YL1130 as a push-pull amplifier driven by a second YL1130 operating as a push-pull frequency trebler. A preceding frequency multiplying stage is provided by a YL1000, a quick-heating v.h.f. power pentode also designed for application as a power amplifier or frequency multiplier in portable or mobile equipment.

To provide an aerial feeder power of 6W at a frequency of 480Mc/s, the circuit requires an input drive power of 0·5 to 0·75W at f = 53·4Mc/s. This requirement was met in the laboratory-produced equipment on which this article is based, by the use of a further YL1000 operating in a crystal-controlled electron-coupled oscillator/trebler circuit. However, as this system would not be a satisfactory commercial proposition, the design of this first stage is omitted from both the circuit diagram and descriptive text of this article, and the availability of a suitable drive source assumed.

The operating conditions for this transmitter are suitable only for frequency modulation. With adjustment to the final stage (see published data for the YL1130) the transmitter can be adapted for amplitude modulation with a carrier power output of 5W.

CIRCUIT DESCRIPTION

A diagrammatic form of the transmitter circuit described is given as Fig. 1, and the inductors detailed in Fig. 2.

YL1000 Frequency Trebler

With the input drive power of 0·5 to 0·75W at a frequency of 53·4Mc/s capacitively coupled to the grid circuit of the YL1000, the centre-tapped anode circuit inductance provides a balanced push-pull output of approximately 1W at f = 160·2Mc/s. The balancing capacitor (C_5 in the diagram) is set to approximately 5pF to simulate the output capacitance and strays of the YL1000.

YL1130 Frequency Trebler

The output from the previous stage is inductively coupled to the grid circuit of the YL1130. The anode circuit, tuned to the third harmonic of the 160·2Mc/s input, provides a balanced output at f = 480Mc/s for the drive of the final stage.

YL1130 Push-pull Amplifier

Further inductive coupling is employed between the anode output circuit of the previous stage and the grid input circuit of the amplifier. A metal plate is mounted across the amplifier valve base to screen its grid circuit from its anode circuit and thus prevent the possibility of instability caused by capacitive or inductive interaction between the circuits.

TUNING AND COUPLING CONSIDERATIONS

Although the theoretical design of such transmitter circuits as this is relatively simple, in practice it is usually difficult to achieve maximum efficiency as difficulties are generally met in meeting the requirements of both drive power and load power. A power output of 6W at 480Mc/s is easily obtained from the YL1130 using a quarter-wave line with shorting-bar tuning, but such a method is not suitable for a commercial equipment.

Quarter-wave Line with Butterfly Tuning Capacitor at Anode End of Line

Experiments have been made to capacitively tune a quarter-wave line at the anode end, but even using the smallest commercially available component the minimum capacitance value is too high, and the resulting losses render the circuit inefficient. Further, the shortened line length creates mechanical problems in that adequate coupling to the next stage is difficult. Attempts were made to use lower impedance lines (physically longer), but the ensuing problems emphasised the unsuitability of such a circuit.

Half-wave Line with Butterfly Tuning Capacitor at Remote End

A half-wave line circuit provides, in effect, two quarter-wave lines back-to-back with a high-voltage open end where a tuning capacitor is most effective. If the chosen tuning capacitor is of similar value to the push-pull output capacitance of the valve, the resultant line length is twice that normally used for the quarter-wave line and shorting bar method, but is considerably less efficient, with a marked

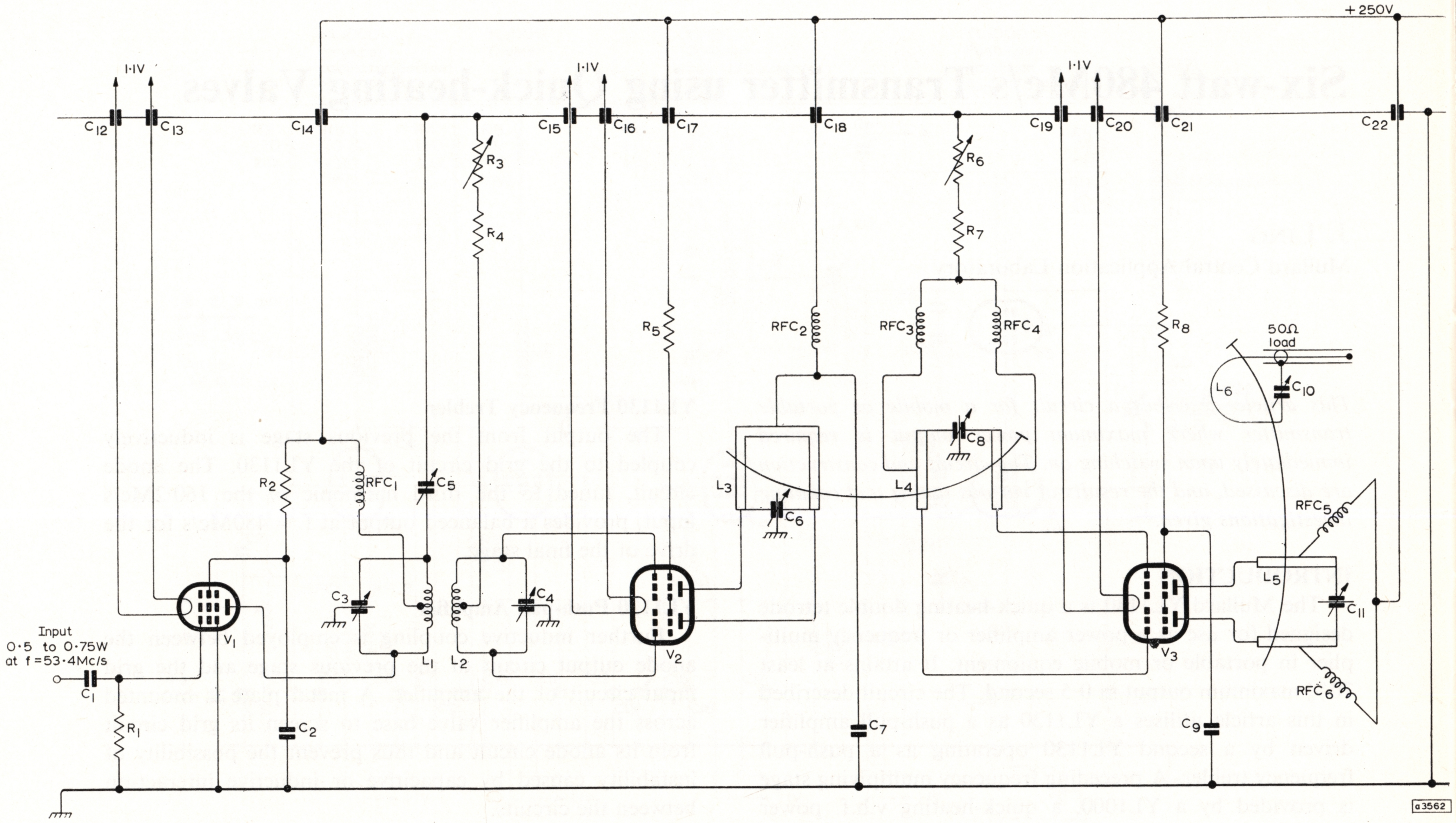

+250V
1·1V
1·1V
1·1V
C12
C13
C14
C15
C16
C17
C18
C19
C20
C21
C22
R3
R4
R5
R6
R7
R8
RFC1
RFC2
RFC3
RFC4
RFC5
RFC6
C5
C6
C8
C10
C11
L3
L4
L5
L6
50Ω load
R2
C3
C4
L1
L2
V1
V2
V3
Input
0·5 to 0·75W
at f=53·4Mc/s
C1
R1
C2
C7
C9
a3562

Fig. 1—Circuit diagram. The cathode return to earth, for all the valves, is provided in the heat er supply circuit. If a transformer is used then it should have a centre-tap and this should be earthed.

COMPONENTS

Inductors

	Type
L_1 to L_6	Detailed in Fig. 2
RFC_1	Painton miniature type 200/220
RFC_2 to RFC_6	Resonant at 480Mc/s

Resistors

All resistors are fixed value ±10%, carbon composition and rated at ¼W unless otherwise stated.

	Value (Ω)	Type	Rating
R_1	82 k		
R_2	82 k		
R_3	100 k	Variable, linear	1W
R_4	22 k		
R_5	100		
R_6	50 k	Variable, linear	1W
R_7	6·8k		
R_8	100		

Capacitors

All capacitors are 250V working
All fixed capacitors have a tolerance of ±20%

	Value (pF)	Type
C_1	33	fixed, ceramic
C_2	4700	fixed, ceramic
C_3	2·5	air-spaced butterfly
C_4	2·5	air-spaced butterfly
C_5	2 to 10	concentric trimmer
C_6	See Fig 6	
C_7	470	fixed, ceramic
C_8	4	air-spaced butterfly
C_9	470	fixed, ceramic
C_{10}	2 to 10	concentric trimmer
C_{11}	4	air-spaced butterfly
C_{12} to C_{22}	1000	lead-through
C_{23} to C_{29}	1500	fixed, ceramic

Valves

	Type
V_1	Mullard YL1000
V_2	Mullard YL1130
V_3	Mullard YL1130

decrease in load power, from 6 to 5W being a typical value. However, if the tuning capacitance value is increased and the line length correspondingly shortened, the circuit losses decrease and a performance practically equal to the quarter-wave line and shorting bar method is obtained. Although, as stated, a reduction in line length is unavoidable, the resultant line length is still sufficient for mechanical coupling without difficulty. These considerations show that such a tuning method would be both suitable and practicable.

Series Tuned Loops

The drive power requirement for the YL1130 used in the amplifier stage is at its lowest when a quarter-wave line is used. Unfortunately, the tuning point of a quarter-wave line at 480Mc/s when loaded by the valve input capacitance is at the base pins of the valveholder, and is thus inconvenient for coupling and tuning. The introduction of series tuned loops serves to increase the circuit length, but the voltage losses across even these small values of capacitance result in a substantial increase in the drive power requirement.

However, a solution can once more be obtained by using a half-wave line tuned at the remote end.

Three-quarter-wave Line

The use of a three-quarter-wave line is not favoured because of its excessive physical size. Folding the line into a smaller space results in additional losses.

TUNING ADJUSTMENT ACCESSIBILITY

Although the capacitor-tuned shortened half-wave line method proves satisfactory for both the amplifier and penultimate trebler stage, some difficulty may be encountered with the mechanical arrangement for optimum coupling in order to introduce chassis-top tuning. This problem arose with the laboratory constructed transmitter discussed here, and the solution arrived at was to return to tuning by the quarter-wave line with capacitive tuning at the anode end method for the trebler stage. As already stated, the main difficulty encountered with this method is the non-availability of a suitable commercial capacitor. This problem was solved by a simple modification to a standard Mullard trimmer to give a small sized balanced trimmer with a low minimum capacitance.

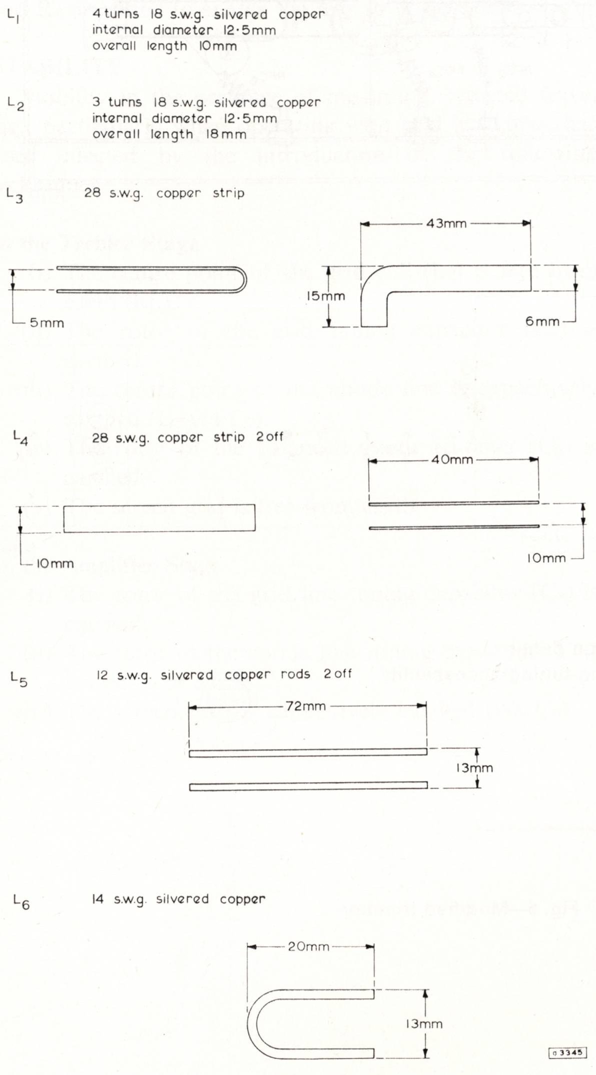

L1
4 turns 18 s.w.g. silvered copper
internal diameter 12·5mm
overall length 10mm
L2
3 turns 18 s.w.g. silvered copper
internal diameter 12·5mm
overall length 18mm
L3
28 s.w.g. copper strip
43mm
15mm
5mm
6mm
L4
28 s.w.g. copper strip 2off
40mm
10mm
10mm
L5
12 s.w.g. silvered copper rods 2off
72mm
13mm
L6
14 s.w.g. silvered copper
20mm
13mm
a3345

Fig. 2—Inductor details

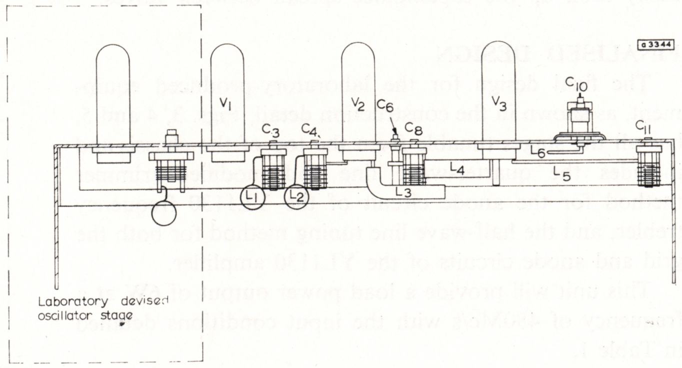

a3344
V1
V2
V3
C3
C4
C6
C8
C10
C11
L1
L2
L3
L4
L5
L6
Laboratory devised oscillator stage

Fig. 3—Construction detail
Side view of chassis showing inductor positioning

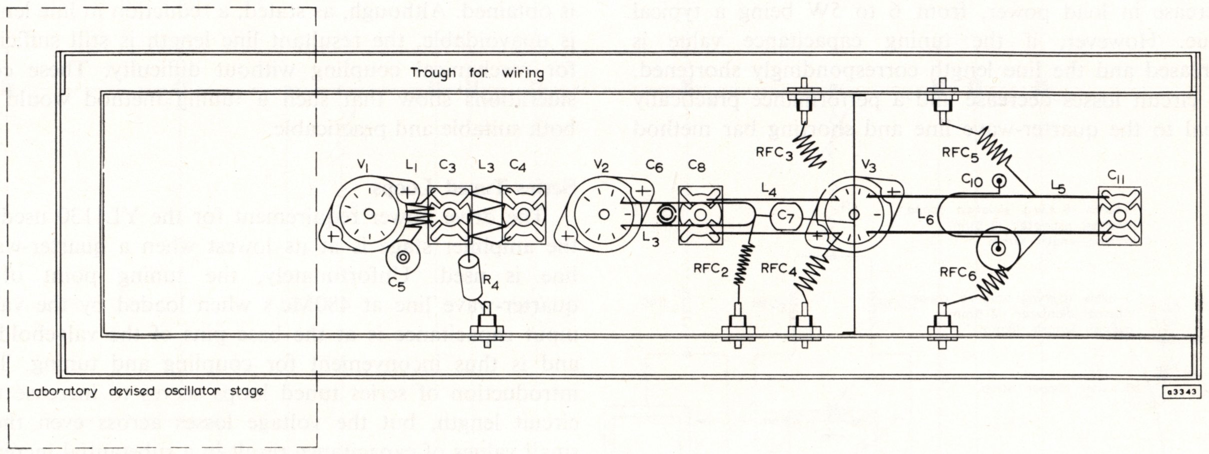


Fig. 4—Construction detail
Underside view of chassis showing inductor positioning

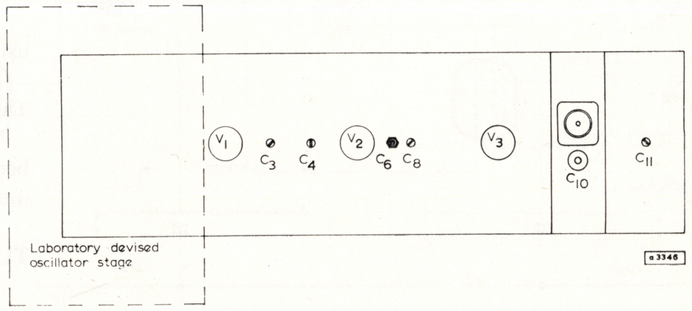


Fig. 5—Construction detail
Topside view of chassis showing tuning accessibility

Modification to Mullard C004EA/12E Trimmer

Fig. 6 details the modification. The external tube 'A', which forms the outer plate, is split longitudinally so that two semi-cylindrical sections are formed. The edges of these sections are filed back so that when reassembled two gaps, each of 1mm width, are formed 180° apart. Reassembly of the item consists of placing the 6mm inside diameter collar 'B' over the two sections of the split plate so that they are retained, equidistant as detailed in the previous sentence, on the ceramic tube. The original threaded plunger is then replaced.

This modified component proved to be satisfactory when used with the quarter-wave line method for tuning the YL1130 trebler anode circuit. It provided an adequate safety margin over the drive power requirements and easily took up the capacitance spread between valves.

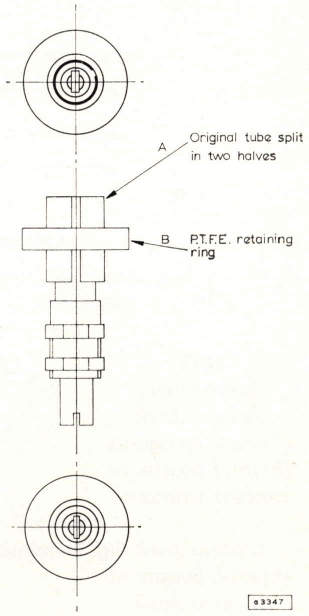


Fig. 6—Modified trimmer

FINALISED DESIGN

The final design for the laboratory-produced equipment, as shown in the construction detail, Figs. 3, 4 and 5, has all the stages tunable from the top of the chassis and includes the quarter-wave line and modified trimmer method for the anode circuit of the YL1130 frequency trebler, and the half-wave line tuning method for both the grid and anode circuits of the YL1130 amplifier.

This unit will provide a load power output of 6W at a frequency of 480Mc/s with the input conditions detailed in Table 1.

TABLE 1
Input Conditions

	V_a (V)	I_a (mA)	V_{g2} (V)	I_{g2} (mA)	V_{g1} (V)	I_{g1} (mA)	R_{g1} (kΩ)
YL1000 trebler	175	27	95	1	−82	1	82
YL1130 trebler	175	45	175	6	−68	1	68*
YL1130 amplifier	175	80	175	11	−22	2·5	8·8*

*R_{g1} adjustable

ADJUSTABLE GRID RESISTORS

Owing to the spread of grid currents between valves both in trebler and amplifier operation, coupled with the fact that the YL1130 has a high mutual conductance, a fixed value of grid bias resistor is likely to cause some variation of input current—and therefore of output power—valve to valve. For these reasons variable resistors R_3 and R_6 are fitted to enable the bias to be adjusted.

STABILITY

Stability in the presence of mistuning, reduced drive, etc., particularly when operating with grid leak bias, has been effected by the introduction of the following techniques:

In the Trebler Stage

(i) The centre point of the grid coil (L_2) is free from earth (r.f.).
(ii) The rotor of the grid tuning capacitor (C_4) is earthed.
(iii) The centre point of the anode line is capacitively earthed (L_3 via C_7).
(iv) The rotor of the balanced anode trimmer (C_6) is earthed.
(v) The screen grid is free from earth.

In the Amplifier Stage

(i) The rotor of the grid line tuning capacitor (C_8) is earthed.
(ii) The rotor of the anode line tuning capacitor (C_{11}) is free from earth.
(iii) The screen grid is capacitively earthed (via C_9).

ADDITIONAL DECOUPLING

Seven 1500pF ceramic capacitors of small physical size are used to provide additional filament decoupling in the amplifier and trebler stages. These capacitors, not shown in the circuit diagram or construction details but allocated circuit reference numbers C_{23} to C_{29} in the components list, are mounted with the shortest leads possible between the filament pins and the nearest earth points.

Four of the capacitors are used in the amplifier stage from pins 2, 3, 7 and 8 to earth, and three in the trebler stage; two from pins 3 and 8 to earth and the third between pins 2 and 7.

POWER CONSUMPTION

The maximum permitted voltage for the anode and screen grid of the YL1130 at 500Mc/s is 200V. The transmitter is designed for a nominal h.t. of 175V at nominal battery voltage feeding the converter. This allows for normal operation with the battery voltage increased by 15% when 'on charge'.

The power lines required for V_1 to V_3 are:

h.t. 175V, 170mA
l.t. 1·1V, 7·1A

In order to reduce hum modulation to a minimum, the filaments should be heated by a balanced supply utilising an earthed centre-tap or other suitable circuit.